Broad Band Data Analysis
Investigation of Skew on Differential High Speed Links

Jie Ji

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Summary

Skew in telecommunication normally means the difference in arrival time of bits transmitted at the same time in differential transmission. As an increasing of transmission data bit rate and more importantly, a data and clock signal rise time of become faster, digital system interconnects became behaving as transmission line. The high speed signals become microwave in nature. The problem is that modern digital designs and verifications require knowledge that has formerly not been needed for a data bit rate of below than 100Mbit but also at the higher frequency range as 5 to 15GHz, however, most references on the necessary subjects are too abstract to be immediately applicable to the skew. For this reason a new method to investigate the skew were introduced, and with which, test board were measured. Since the test boards are made in devise material, and lines on the boards are configured out in distinct structures. In this paper, several methods were applied to find out the skew, and by comparing the results, it could be found that how factors affect the skew, not only the material factor, but some manufactory reason.
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1 Introduction

1.1 Work background

The balanced signal has been used for signal transmissions in telecommunication equipment instead of single ended configuration since 1980s [4]. The advantage is mainly based on immunity and grounding aspect from system point of view [8]. Skew in telecommunication normally means the difference in arrival time of bits transmitted at the same time in parallel transmission [6]. And this normally presents as common mode signal in balance transmission, which is deleterious to the signal communication, and could destroy the board if accumulated much [32]. Therefore the most important role of a balanced line interconnect is to transmit signal from one point to another with acceptable skew. Hence, measuring the skew and investigating the factor to them become pre-requisite.

1.2 Measurement methods used for skew

In this decade measurement for skew has been developing. Time domain reflection and transmission have been adopted to measure the skew since 1990s until now as the most directly methods to investigate the time difference in time domain [12]. And the advantage appears in its instant response for resistance by inject perfect compensate microwave signal. At the meanwhile, to check skew in frequency domain, mostly VNA vector network analyzers have been used with SOLT short open load transmission calibration with the advantage of the convenient of using and low cost of universal calibration kits in the recent decade [13]. Both of the method mentioned before are working accurately in frequency range below 5GHz. However, as the requirement of the LTE architecture, the frequency more than 5GHz was required [14] [29] [31]. And in these frequency skew measurement becomes complicate because interfere of the test fixture could mask the skew [3] [26] [48]. In these years, Simulation to cancel interfere of the test fixture has been processing yet have obviously breakthrough, because of the bottleneck to model asymmetry of material [16] [18]. And as an increasing of transmission data bit rate and more importantly, a data and clock signal rise time of become faster, digital system interconnects became behaving as transmission line. The high speed signals become microwave in nature [39]. The problem is that modern digital designs and verifications require knowledge that has formerly not been needed for a data bit rate of below than 100Mbit but also at the higher frequency range as 5 to 15GHz, however, most references on the necessary subjects are too abstract to be immediately applicable to the skew[1][29][41].
1.3 Problem statement

Since the transmission should be run in high frequency range as the new generation communication system required. Transmission quality should be controlled in a precise standard. Therefore, skew measurement in high frequency range and its relative factor analyze become a problem starved for improvement. In this project, we investigate the skew of hundreds signal circuits with TDR, TDT, group delay check, and S parameter measurement in the test part [7] [11] [29] [43]. And as a result that only S parameter measurement with TRL calibration is able to remove the effect of the test fixture [15] [19] [21] [23] [35] [47] [52]. As the result of TRL calibration, new phenomenon was found and it declares some results which have misunderstood for many years, which will be present in details in this paper. And in the analysis part, we try to conclude the factors affecting the skew, both in material and manufactory aspects. Normally say, different length between the multi-conductor structures will make skew, however, in the higher frequency range, even any unbalance device would produce skew [11][28][34][45][50]. And the unbalance cannot be avoided for the material changing, manufacture inaccurate and even some stabs in via holes [49]. So we will measure different boards with diverse material, on which tens of distinct pairs of line located. And a new model has been built successfully in this project to calculate the material data from S parameters [5][8][9][19][23]. And disparate typical manufactory problem are also preformed on these two test boards. So that how they effects skew was investigated.

1.4 Future work

Through this work, some factors effecting skew was found, however, to convince them, we need to measure new test boards with TRL calibration, because it is the only one in this stage could cancel the interfere of the test fixture.
2 Theoretical analysis and mathematical modeling

2.1 Why differential Mode

The advantage of differential signal is higher noise immunity achieved by eliminating common mode influences picked up by the environment. Provided its conductive traces are close enough together to be exposed to nearly the same environmental influences, the differential transmission line is desensitized to voltage drift because the induced noise in each line rise and fall in tandem in the complementary signals and cancel each other. The inherently superior noise immunity of differential circuits allows signal level to be reduced, with consequent improvements in switching speed, power dissipation and noise. Another advantage lies in the area of ground bounce. Since differential pairs always sink and source the same amount of current. There is no bounce. And in theory no bounce means no crosstalk. And since the currents are complementary there is little net magnetic flux from a differential pair, so EMI is dramatically reduced as well.

2.2 What is Skew

The classical view of cable skew in high-speed differential signaling is that of two independent signal paths of differing propagation delay. The following diagram accurately presents this view. Unfortunately reality is not as simplistic as this.

The figure below shows the skew measurement. Both rising edges of the two signal trace are intended to change state simultaneously as shown on the left side. When this doesn’t occur at the time offset between the switching states is referred to as the skew. And this displayed on the right side.
The midpoint or switching point of each edge is typically the measurement reference point. And this paper the former one was picked.

Any negative effects of the display are solely due to the components. When applied to the output drivers of a component with sharp and clean edges, the measurement is rather straightforward.

This imbalance can also be introduced by interconnect. In this case, measurement requires the near end of the interconnection to be driven by lab quality signals, and at the mean while, the offset between the state transitions measured at the far end is the skew. The signal injected may have clean crisp edges but the signal observed at the far end frequently doesn’t. And further more, when signals were applied on interconnect, skew is more difficult to measure for the effect of the connector. And in this paper a method to calibrate away the connector would be introduced.

And in nowadays signal transmission, differential lines are used. Each differential pair consists of a positive and a negative leg, which are assumed to be perfect complements of each other. And this assumption is to the health and design objectives of the differential signal interface. In this type of application imbalance is frequently attributed as a contributor to some system level problems, such as inadequate signaling margins and excessive electromagnetic emissions.

2.3 Mathemetic Analysis and Modeling

Factors affecting the skew are various, classed as material factors and manufactory factors. And as manufactory factors always located in a part of the board, at the mean while the material factor count all over the plat, so it may be the big problem and easy to be found and on the contrary, the manufactory factor is the subtle one and hard to be seen
2.3.1 Material

A way of describing each loss term, is the resistance and conductance per length of the lossy line. These two terms factor into the attenuation of the line.

$$\alpha = \frac{1}{2} \left( \frac{R_L}{Z_0} + G_L Z_0 \right) \text{ [nepers/length]} \quad (1)$$

$$= 4.34 \left( \frac{R_L}{Z_0} + G_L Z_0 \right) \text{ [dB/length]}$$

And the attenuation term reduces to a simple form that includes the attenuation from the resistive losses and the dielectric losses, which were presented as

$$\alpha = \alpha_{\text{dielectric}} + \alpha_{\text{conduct}} \quad (2)$$

And

$$\alpha_{\text{conductor}} = 4.34 \times \frac{R_L}{Z_0} \quad (3)$$

$$\alpha_{\text{dielectric}} = 4.34 \times G_L Z_0 \quad (4)$$

Where

$$G_L = \omega \tan \delta C_L \quad (5)$$

$$R_L = R_{AC} + R_{DC} \quad (6)$$

And where $R_{AC} = \frac{\rho}{W} \left( \frac{1}{t} \right)$, $R_{DC} = \frac{\rho}{W} \left( \frac{0.75 \sqrt{f}}{63 \mu} \right)$. [9]

In the later section, how the equations above were derived will be discussed in details.

As the equations above schemed that at low frequency, the conductor loss dominate the loss factor, and if the frequency goes higher as 5 to 20GHz, the dielectric factor became the mainly one. And there is no change in geometry which will decrease the dielectric loss. It is simply a material selection issue. If dielectric losses dominate, and the attenuation must be decreased, the only way of doing this is by changing the PCB laminate material.

And according to the equation 12, that less $\tan \delta$ brought less dielectric loss. So whether this conclusion is correct or not, the lab contrast in the following section would make sure.

With such reason, to find out which material brought less loss, the section below will focused on such points.

First, a model will be introduced, with which the RCGL parameters of the transmission line and the dielectric constants of the material could be draft out. In addition, as the previous section the result of different material will be discussed with mixed-mode to show how the material effects the transmission and conversion.
2.3.1.1 Extract Propagation constant from S-parameters

For a transmission line, basic parameter for a line can be summarized as RCLG parameter in a ladder circuit as follow: [3]

![Figure 3: the ladder circuit representation of transmission line](image)

\[ γ = \alpha + j\beta \]

\[ Z_S = \frac{R + j\omega L}{G + j\omega C} \]

\[ \gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \]  

\[ \gamma \] is complex. The real part of the propagation constant, usually termed \( \alpha \), is the attenuation per length. The imaginary part usually termed \( \beta \), is related to the wave velocity.

Solving equations of (7) and (8), RCLG for the single ended line can be derived as:

\[ R = \text{Re}(jZ_S) \]
\[ L = \text{Im}(jZ_S)/\omega \]
\[ G = \text{Re}(\gamma / Z_S) \]
\[ C = \text{Im}(\gamma / Z_S)/\omega \]

Combining (9) to equation (7), RCLG can be figured out with S-parameters in the following specification.

A uniform transmission line as a two port network was schemed as bellow,[5]

![Figure 4: uniform transmission line](image)
The $S$ parameters are defined for a two-port network in terms of the total voltages and currents as shown in Figure 5 and the following: [3]

$$
\begin{align*}
V_1^+ &= S_{11}V_1^- + S_{12}V_2^- \\
V_2^- &= S_{21}V_1^+ + S_{22}V_2^-
\end{align*}
$$

(10)

Figure 5: A two-port network

The input and output voltage and current are showed as follow:

$$
\begin{align*}
V_1 &= V_2 \cosh \gamma l + I_2 Z_0 \sinh \gamma l \\
I_1 &= \frac{V_2}{Z_0} \sinh \gamma l + I_2 \cosh \gamma l
\end{align*}
$$

(11)

Where $Z_0$ is characteristic impedance, $l$ is the length of the transmission line, and $\gamma$ is the prorogation constant.

The element of scattering matrix is given by

$$
S_{ij} = \frac{V_j^-}{V_i^+}
$$

(12)

Introduce the reflection coefficient between a medium and the vacuum $\Gamma^-$, the signal flow graph is shown as: [12]
Then the single end reflection and transmission S-parameters of the
electromagnetic wave through a transmission line can be shown to be [1]

\[
S_{11} = \frac{V_1^-}{V_1^+} = \frac{(1-e^{-j\theta})\Gamma}{1-\Gamma^2e^{j\theta}} \quad \text{and} \quad S_{21} = \frac{V_2^-}{V_1^+} = \frac{(1-\Gamma^2)e^{j\theta}}{1-\Gamma^2e^{j\theta}} \quad (13)
\]

Then \( \Gamma \) can be solved based on the equation (13) as:

\[
\Gamma = \frac{1+S_{11}^2-S_{21}^2 \pm \sqrt{S_{21}^4 + S_{11}^4 + 1-2S_{21}^2S_{11}^2 -2S_{21}^2 -2S_{11}^2}}{2S_{11}} \quad (14)
\]

Based on the measurement result showed bellow, when `'+' was picked to
calculate the \( \Gamma \), it would become close to 0 when frequency goes high. And `'-' option is not.
For $\Gamma$ is the reflection coefficient, which cannot be equal to 0, ‘−’ was selected. Then $Z_0$ and $\gamma$ were expressed as:

$$\gamma = -\frac{1}{\Gamma} \ln\left(\frac{1 - S_{11}}{S_{21}}\right)$$  \hfill (15)$$

$$Z_0 = 50 \times \frac{1 + \Gamma}{1 - \Gamma}$$  \hfill (16)$$

So that the single end propagation constant and the characteristic impedance can be configured out with S parameters.

The practice situation, however, the transmission line is dealing differential mode instead of single end. To obtain the propagation parameter for differential mode, the single end parameters CGRL should be transform to differential mode parameters $C_d\ G_d\ L_d\ R$.

And any single segment of differential mode transmission line could be schemed as

- a. differential signal with ground
- b. differential signal with virtual reference

Figure 8: The ladder circuit representation of differential mode transmission line
where \( C_{d1} = C_1 + 2C_m, C_{d2} = C_2 + 2C_m, G_{d1} = G_1 + 2G_m, G_{d2} = G_2 + 2G_m, \)
\( L_{d1} = L_1 - L_m, L_{d2} = L_2 - L_m. \)
Obtaining these parameters, such \( C_d, G_d, L_d, R \) could be constitute a uncoupled circuit for differential signals

\[
\begin{align*}
\text{Figure 9: the ladder circuit representation of differential mode presented with uncoupled circuit}
\end{align*}
\]

Such figure is a single end mode, but the parameters in it is differential,
\[
\begin{align*}
C &= C_{d1} + C_{d2} \\
L &= L_{d1} + L_{d2} \\
G &= G_{d1} + G_{d2} \\
R &= L_1 + L_2 \\
\end{align*}
\]
Since it is a single end model, the propagation constant could be drawing out as the following section discussed.

### 2.3.1.2 Modeling building to Extract loss factor from propagation constant

To model attenuation in a single end transmission line, we need to add the lossy elements of the series resistance per length and the shunt conductance per length to the typical lossless equivalent circuit model for a transmission line. The model is shown in the figure 31. The conductance term is really a resistor that represents the AC loss of the dielectric.

Given this single end transmission equivalent circuit model showed in figure 31, it is possible to solve for the propagation of sine waves. What we find is that sine waves will propagate as sine waves, with a single end propagation constant, usually termed \( \gamma \), that is complex. The real part of the propagation constant, usually termed \( \alpha \), is the attenuation per length. The imaginary part usually termed \( \beta \), is related to the wave velocity. [9]

\[
\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}
\]

As low loss approximation that
\[
R_L << \omega L, \quad G_L << \omega C
\]
Then the $\alpha$ and $\beta$ can be calculated as

$$
\beta = \omega \sqrt{L_C} = \frac{\omega}{V}
$$  \hspace{1cm} (19)

$$
\alpha = \frac{1}{2} \left( \frac{R_L}{Z_0} + G_L Z_0 \right) \text{ [nepers/length]} \hspace{1cm} (20)
$$

$$
= 4.34 \left( \frac{R_L}{Z_0} + G_L Z_0 \right) \text{ [dB/length]}
$$

The attenuation term reduces to a simple form that includes the attenuation from the resistive losses and the dielectric losses, which were presented as

$$
\alpha = \alpha_{\text{dielectric}} + \alpha_{\text{conductor}}
$$  \hspace{1cm} (21)

And

$$
\alpha_{\text{conductor}} = 4.34 \times \frac{R_L}{Z_0} \hspace{1cm} (22)
$$

$$
\alpha_{\text{dielectric}} = 4.34 \times G_L Z_0 \hspace{1cm} (23)
$$

Where

$$
G_L = \omega \tan \delta C_L \hspace{1cm} (24)
$$

$$
R_L = R_{AC} + R_{DC} \hspace{1cm} (25)
$$

And where $R_{DC} = A_1 \sqrt{f}$, $R_{AC}$ and $A_1$ are functions of the material and shape of the line. \[9\]

Combining the equations (21) to (25), the attenuation term can be configured out as

$$
\alpha = 4.34 \times \frac{R_{AC}}{Z_0} + 4.34 \times A_1 \cdot \sqrt{f} + 2.3 \sqrt{\varepsilon_{\text{eff}}} \tan \delta \cdot f / c_0
$$

$$
= A_0 + A_1 \sqrt{f} + A_2 f
$$  \hspace{1cm} (26)

Since the propagating constant was extracted from S parameters, the attenuation term of it can be figured out. With matlab curve fitting, $A_0$, $A_1$, $A_2$ should be obtained, then the loss factor $\sqrt{\varepsilon_{\text{eff}}}$ and $\tan \delta$ would be extracted.
2.3.2 Manufactory Factors

2.3.2.1 Different way to Calibration

The material effect is clear to see. Some effect by connector could not cover up them. However, when some manufacture factors were investigated, the effect of connector would be a problem; such i.e. will be showed in the following section. So some other calibration which could eliminate some connector effort was required. And in this section, different calibration will be introduced one of them will be selected to apply in the following measurement.

2.3.2.1.1 Measurement Error and Correction

The ideal properties of a NA, so far being described, are only partly possible to realize [12]. Deviations from ideal behavior are giving rise to measurement errors divided into two groups, stochastic and systematic errors. Examples of stochastic errors are temperature drifts and drifts over time, noise from the signal source and the receiver. They are bad repeatability of circuit. Examples of systematic errors are positive values of directivity of devices for signal separation, mismatch at the test ports, and cross-talk between ports. They are reproducible. Since systematic errors are reproducible, in contrast to stochastic errors, corrections may be performed compensating such errors.

The influence of the stochastic error has to be minimized by proper design of system components. The dominating sources of error are the signal source, the signal- separation unit, and the receiver. Since the stochastic errors are not reproducible and cannot be corrected for. In the real measurement, only system error was correction.

As opposed to stochastic errors, systematic errors are reproducible, thus making them possible to be corrected if known. An error-correction technique is applied for the purpose, where a linear error model is commonly used describing the deviations of the NA from ideal behavior.

Two-way NAs are advantageous since all four s parameters of a two-port can be measured without turning the DUT around. Moreover, full correction for mismatch at the test ports is possible for transmission measurements. Measuring in the forward direction, the one-way error model outlined above may be used. Adding an extra error term, $X_F$, for cross-talk from port 1 to port 2 is also possible. Measurements in the reverse direction are performed analogously. Since matching and attenuation in the test set of the NA usually are changing when the measurement direction is changed, independent error models are achieved for the measurement in the forward and backward directions, thus resulting in 12 error terms, see figure below. Only three receiver channels are requested for
each measurement direction, and, thus, the 12-term correction model is particularly well suited for two-way 3 channel NAs, where the reference channel is disconnected before the signal from the signal source is passing the RF switch.

In the 12-term model, the cross-talk errors are assumed constant. In practice, however, cross-talk errors are often DUT dependent. [12]

2.3.2.1.2 SOLT Fixture Calibration Standard

The 12-term model calibration procedure SOLT is based on a complete one-port calibration of both test ports and a THROUGH measurement, concurrently. As the THROUGH standard is connected, the reflection is also measure in order to determine the reflection coefficient of the opposite port. When isolation is measured both ports have to be matched.
The Slot procedure is using a one-port standard with unknown high reflection, i.e., a REFLECT. A REFLECT may be a SHORT or an OPEN with high attenuation and an offset length specified with an accuracy of a quarter of a wavelength. The length information is requested for the elimination of the sign uncertainty when correction coefficients are calculated. However, it is important that the unknown reflections are exactly equal at the two test ports. For this reason, Slot calibration can only be used for symmetrical connector systems, if the connectors are of the same type at both ports, or if the male connector of the REFLECT is of equal electrical properties as the female connector. The Slot is particularly well suited for transmission line types, which are not facilitating a good OPEN or SHORT. And it made an accurate calibration in a wild band width from 10MHz to 20GHz.

2.3.2.1.3 TRL Calibration Standard

Same with the Slot calibration standard, 12 terms error model were introduced on the THROUGH-REFLECT-LINE calibration.

The TRL procedure is different from the Slot procedure exchanging a LINE for a MATCH as the calibration standard fixing the Smith-chart center. The LINE is the same with the wave guide used on the DUT. In this project, the wave guide is the transmission line on PCB.

A TRL calibration is bandwidth limited due to a singularity which appears if the difference in electrical length between a THROUGH and a LINE is a multiple of a half wave length. The maximum ratio between upper and lower band edges is limited for calibration to 8:1. To manage a larger frequency range in practical applications of the TRL, usually two or more LINEs of different lengths are used, and the LINEs are also the same with the wave guide which will be used on the DUT. A combination with Slot for the lower frequency range is also possible, since well matched loads for those frequencies can be made. If the THROUGH standard is replaced by another LINE, the TRL procedure works.

And the advantage of TRL compare with the Slot presents on the high frequency range. For the Slot calibration only works on the ports, in another word, Slot only calibrates away the error effect from the one port through the network analyzer till another port. So the effect of the connectors is not able to be eliminated. However, the TRL, although the accuracy is less than the SLOT electric calibration, it could calibrates away the influence since some point of a line through one connector, and the analyzer, via another connector till another symmetry point of another line, as the figure showing below.
TRL calibration eliminated the effect of the connector, and made the transmission much better and the reduced the conversion from differential mode to common mode, which means the skew on differential lines, especially in the frequency as 5GHz to 15GHz.

L5_CF2_line1 on Test board A was picked up as i.e.

![Figure 12: difference between the SOLT and TRL calibration standard](image)

According to this picture, with TRL calibration, the insertion loss is about 3 to 6 dB lower, compare to the result with SOLT calibration, and as the frequency goes up, the advantage of the TRL calibration gets more visible. As the TRL calibration, eliminated the effect of the connector, which brought the much insertion loss, and as the frequency goes to higher range, even above 15GHz, the effect of the connector, almost destroyed the transmission.

![Figure 13: Sdd21](image)
And for more important, take the L5_CF2_line1 and L5_CF2_OFFSET_line1 as i.e., with SOLT calibration, the transmission was

![Graph showing transmission with SOLT CAL](image)

Figure 14: the transmission with SOLT CAL

It was obvious that with the SOLT calibration, there is very subtle difference to be found out on these two different structures of line. In another word, the connector effect covered up the difference of the offset line and the standard line. For this reason, TRL calibration was used in the following measurement.
3 Methodology
The measurement of the test board was done in time domain and frequency domain respectively to find the skew with different methods.

![Measurement methods](image)

Figure 15: Measurement methods

3.1 Time Domain

3.1.1 TDR

Time Domain Reflectometry (TDR) has traditionally been used as the key measurement technology for electrical characterization of electronic packages. The step generator produces a positive-going incident wave that is applied to the transmission system under test. The step travels down the transmission line at the velocity of propagation of the line. If the load impedance is equal to the characteristic impedance of the line, no wave is reflected and all that will be seen on the oscilloscope is the incident voltage step recorded as the wave passed the point on the line monitored by the oscilloscope. Any discontinuity on the transmission line would be detected and schemed as a surge on the oscilloscope immediately.

![TDR configuration](image)

Figure 16: Configuration for TDR/TDT measurements

A step signal was injected into the DUT, which is the left side showing on the above figure. And the trace length can be calculated from the time delay on the signal traces, which can be read as follow:
When a pair of step signals was applied at the input port, which consists of a positive and negative leg and are assumed to be perfect complement of each other, the time difference detected was referred as the skew, which can be measured with TDR as follow:

First the lengths of the positive signal trace and the negative signal trace were measured respectively.

Then the time difference of the two traces can be calculated out,

\[ \Delta t = \frac{|t_{\text{negative}} - t_{\text{positive}}|}{2} \quad (27) \]

As the case before, the time difference is 18 ps.
3.1.2 TDT

When the far end of the same transmission line was connected as a receiver, this measurement was called time domain transmission, or TDT. A schematic of this is shown in Figure 1.

When measure one trace only, the transition situation can be read on the screen as followed.

In TDT, something likewise, pair of perfect matched step signal was injected in port 1, and the time difference of the trace can be read on the screen directly.

The following figure shows the response of a transmission line to a step excitation. One end of the interconnect model is driven with a step source model while the other is ideally terminated. The later side is probed and the results are shown below.

![Figure 19: signal at port 2](image)

Expanding the transition region how much skew was read. Estimating the vertical midpoint of an edge that has been subjected to large a level of loss can be rather difficult since the final vertical level of the transition is delayed. In this case, 18ps skew was measured at the midpoint as shown below in figure 8. Note that the amplitude is normalized to the initial value at the load.
However, at a level slightly higher than the midpoint a measurement result of 10ps could be obtain and yet another result when the tester go below the midpoint. And distortions in the shape of the waveform near the midpoint and a slow transition to the final value complicates the measurement. Clearly, this measurement is much easier when the transition edges are straight.

These methods, however, relies some disadvantage. First of all, the step wave were inject via the connectors, the common mode made by the connector cannot be excluded. And more important, the skew defining here is man-made. Such as the reference point is the tester selected, and the start and end point of the signal trace are read artificially as well.

### 3.2 Frequency Domain

#### 3.2.1 Group Delay

Transmission lines are associated with phase shifts that are linearly dependent of the frequency, i.e., phase distortion.\[12\] The slope of the phase-response curve is not constant. Minimum and maximum values of the phase shifts are not contributing with information enough to characterize the response, since the speed of the phase variation is not included. The frequency derivative of the phase response \( \frac{d\phi}{df} \), is called the group delay, and this is a measure of the time required for a signal of a specific frequency
passing the transmission line. A linear phase response is giving a constant value of the group delay, i.e., all frequencies are passing the transmission equally fast. Deviations from a linear phase response, i.e., \( \frac{d \phi}{df} \) is not constant over the frequency range, are resulting in deviations from a constant group delay. Such deviations are causing distortion of the transmission signal. The group delay is often a more accurate description of the phase distortion than the phase response.

And the group delay is relative with the electrical length. A radio signal in vacuum is propagating with the speed of light. In a PCB transmission line, the propagation speed is reduced depending on the relative permittivity, \( \varepsilon \), of the insulating material. The time needed for the signal to pass the transmission line can be measured, and thus, the electric length can be determined. The speed of the signal in the transmission line is calculated by:

\[
\nu = \frac{c}{\sqrt{\varepsilon}} \text{ (m/s)}
\]

Where \( c \) is the speed of light in vacuum, and \( \varepsilon \) is the relative permittivity of the insulating material.

The speed multiplied of the measured delay is giving the electric length. And the electric length is calculated with:

\[
l = \nu t = \frac{ct}{\sqrt{\varepsilon}} \text{ (m)}
\]

where \( t \) is the measured delay in the transmission line.

So the electric length could be calculated indirectly based on the group delay. And inferentially, with the same physically length, the offset in time delay with each pair of line is referred as the skew.

![Figure 21: the measurement of group delay](image-url)
Figure 22: the group delay of the pair of line

And relate to this measurement showed in figure 10, the skew is

$$\text{skew} = t_A - t_B$$ (30)

However, this method may be unpractical. For the reason that every time when the group delay of one trace was measured, the other trance was terminated with 50ohm load, so the single ended signal was actually measured instead of the differential mode signal. For the field of these two kinds of transmission are totally different as the figure shows below, so the group delay result could only be an reference instead of a judgment.

A, differential mode

B, common mode

Figure 23: common mode field and the differential mode field

In a word, characterization of differential signaling imbalance by skew is too simplistic. Skew only addresses the offset of two signals in time. The real problem can not be described by just the parameters of phase delay and amplitude but the variation of both with respect to frequency. Skew analysis applies a single dimensional view to a multi-dimensional problem. To detect some subtle problem in 5 to 20GHz, such methods above is too abstract to be applied for the reason we discussed before, so the result of which can only be a reference instead of the judgment. And to get a more accurate result, another method which is so called mix-mode will be
introduced in the next section. With measure the S parameter of the coupled line and analyze the transmission of differential mode and conversion loss of differential mode to common mode how unbalance the device is will be known, and some information of the skew could be defined.

3.2.2 Mixed Mode

3.2.2.1 Terms in Mixed Mode

Balanced circuits have been used for many years because of their desirable performance characteristics. They have been mostly used in lower frequency analog circuitry and digital devices, and much less so in RF and microwave applications.

One benefit of differential circuits is that they have good immunity from many sources of noise such as that from power supplies, adjacent circuitry, and other external sources that are coupled either electrically or electromagnetically. These noise sources tend to be coupled in the common-mode, and therefore cancel in differential mode.

In a balanced device, two terminals constitute a single port. Each balanced port will support both a common-mode and a differential-mode signal. This performance is described using mixed-mode s-parameters.

A mixed-mode s-matrix can be organized in a similar way to the single-ended s-matrix, where each column represents a different stimulus condition, and each row represents a different response condition.

Unlike the single-ended example, though, in the mixed-mode s-matrix the port are not only considered, but also the mode of the signal at each port.

The naming convention for the mixed-mode s-parameters must include mode information as well as port information. Therefore, the first two subscripts describe the mode of the response and stimulus, respectively, and the next two subscripts describe the ports of the response and stimulus.

The mixed-mode matrix fully describes the linear performance of a balanced two-port network. To describe the information contained in the mixed-mode s-matrix, it is helpful to examine each of its four modes of operation independently by dividing this matrix into four quadrants.

As showing in the quadrant in the upper left corner of the mixed-mode s-matrix describes the performance with a differential stimulus and differential response, which describes fundamental performance in pure differential-mode operation.
And for a device with two balanced ports, the quadrant in the lower right corner of the mixed-mode s-matrix describes the performance with a common-mode stimulus and a common-mode response. When the performance of the device is isolated to this specific mode, these four parameters describe the input and output reflections, and the forward and reverse transmissions.

\[
\begin{bmatrix}
S_{dd1} & S_{dd2} & S_{dc1} & S_{dc2} \\
S_{cd1} & S_{cd2} & S_{cc1} & S_{cc2} \\
\end{bmatrix}
\]

Figure 25: CC quadrant

And the parameters in the lower left corner describes the common-mode response of a device to a differential stimulus. As with the other modes, there are reflection parameters on each port, and transmission parameters in each direction.

\[
\begin{bmatrix}
S_{dd1} & S_{dd2} & S_{dc1} & S_{dc2} \\
S_{cd1} & S_{cd2} & S_{cc1} & S_{cc2} \\
\end{bmatrix}
\]

Figure 26: CD quadrant

At last but not the least, the parameters in the upper right corner describes the differential response of a device to a common-mode stimulus. Again, there are reflection parameters on each port, and transmission parameters in each direction.

\[
\begin{bmatrix}
S_{dd1} & S_{dd2} & S_{dc1} & S_{dc2} \\
S_{cd1} & S_{cd2} & S_{cc1} & S_{cc2} \\
\end{bmatrix}
\]

Figure 27: DC quadrant
Both the DC quadrant and the CD quadrant describes conversion, the only difference is the DC presents the differential mode to common mode, and the other shows the common mode to differential mode. In an ideal balanced device that is perfectly symmetrical, there will be no conversion from differential mode to common mode. In that case, all of these terms will be equal to zero. As the device becomes asymmetrical, these terms become larger. Therefore, the mode conversion terms provide a measure of device symmetry.

All of the performance benefits of differential circuits assume that the device is symmetrical. The benefits become diminished as the device becomes more asymmetrical.

Where differential to common mode conversion is related to the generation of EMI in a balanced device, the common to differential terms are related to the susceptibility of a device to EMI. Common mode noise, for example, can become converted to differential mode and degrade the signal-to-noise ratio of the system, which presents on time domain is so called skew.

As differential mode was used normally in signal transition, insertion loss of a pair of couple line present on s parameter is Sdd21. And insertion loss profiles of both samples are shown in figure 16 one of which with greater skew has more insertion loss than the other of which with less skew, throughout the frequency range displayed. And as the frequency goes higher, the insertion loss of the couple line with larger skew is increasing. In this i.e. that the insertion loss is approximately 10 dB difference at 5GHz but increased as about 20dB at 15GHz.

![Figure 28: difference on insertion loss with variable skews](image)

The insertion loss difference is likely the dominant factor in the reduced eye height and not the skew. However, focusing on the amplitude attenuation at one frequency may prove to be as misleading as skew. Nuances in the amplitude and phase response are captured by the s-parameters but may require simulations of
the eye diagrams to expose the real effect on signal margins of NRZ data streams. Additionally, it could be envisioned that other frequency dependent parameters such as data rate or bit pattern can interact with the interconnect to create a more complicated relationship. Modeling tools are necessary to accurately gauge the health of a system and evaluate the signal margin of NRZ data streams. Of course, performing actual eye diagram measurements will hopefully produce similar results.

As described above, Scd21 present the forward transmission of a common mode response to a differential mode stimulus, which yields the EMI. And here it could be found in the figure below that the signal with less skew made lower conversion loss. At the 10GHz, there is 24 dB conversion loss difference between the two signals traces.

![Figure 29: difference of conversion loss of variable signals](image)

Simulated skew can then be compared to the common mode to differential conversion, which should be 0 at an ideal condition, however, in practice situation, there is no perfect device. As showing above, the signal with smaller skew made less conversion than the one with larger skew.

Skew is made by the asymmetry of the device, so it should be independent of the length. Theoretically, lines in the same type with different length should present the similar conversion in mixed-mode. However, the term Scd21, was effected by Sdd21 and Scc21, that means, the real conversion loss may be larger than the value measured, for the reason following.

For one thing, Sdd21 is getting smaller as the frequency goes up higher.

Take line 3 on the test board A (see test sample), which has a longest length as an i.e.,
the conversion loss adown on the figure instead of upgrade when the frequency goes higher, and this didn’t match the theory knowledge.

For one thing, in the differential mode transmission, it could be found that at the higher frequency range only a subtle signal transmitted on the lines, so that the common mode response of a device to the differential stimulus is small as well.

For another thing, the common mode transmission also decreases as the frequency increases,
It’s also goes down for the propagating.
In addition, the common mode was produced in every single segment of the transmission line, and what we measured on the port 2 is the sum of the common mode made from the nearest segment and some propagating result from the former segments. As the figure showed below,

![Diagram showing common mode transmission](image)

**Figure 32: the common mode transmission**

Therefore, the real conversion loss may be larger than the value measured, for the transmission line lost a large part of the differential mode and common mode signal.
Since such was discussed before, for some longer line which has a large insertion loss, the conversion loss may only be a sort of reference instead of a judgment to its skew.
But anyway, refer to some other line with short or middle length, the common mode to differential mode conversion, which will be found out with the mixed-
mode, would present any unbalance of the coupled line accurately. And this judgment is the best choice to find out the skew in high frequency range.

### 3.2.2.2 Mixed Mode Calculation__ Single end to mixed mode

Mixed mode could be calculated out from single end measurement. In single ended device, the signal paths are referenced to a common ground potential as illustrated in figure 22.

![Single end circuit diagram](image)

After measured the S parameter of such coupled line, a 4×4 matrix is obtained as

\[
\begin{bmatrix}
S_{11} & S_{12} & S_{13} & S_{14} \\
S_{21} & S_{22} & S_{23} & S_{24} \\
S_{31} & S_{32} & S_{33} & S_{34} \\
S_{41} & S_{42} & S_{43} & S_{44}
\end{bmatrix}
\]  

(31)

For balanced circuits, by comparison, are constructed with a pair of signal paths where each side of the pair is a mirror image of the other, as illustrated in figure 23. Ideally, it is perfectly symmetrical. When a balanced device is driven such that the two sides of the device are 180 degree out of phase, the relative signals along a perfectly symmetrical device will be of equal amplitude and opposite phase. This creates a “virtual ground” along the axis of symmetry.[5]
One of the important issues for a balanced circuit is that the common mode currents produce radiated disturbance. In an ideal situation, perfectly balanced differential mode currents flowing on a pair path are equal in amplitude and opposite in phase. The fields radiated by the two closely spaced currents cancel out, and such perfectly balanced conductor pair does not create significant signal levels of unwanted radiated disturbances. From SI points of view, qualities and characteristics of a differential mode signal should be addressed; for EMI, the behavior of a common mode signal should be emphasized.

In general, each port can support the propagation of differential mode and common mode waves. Figure 24 shows a conceptual diagram of mixed mode two-port measurement system.

Applied voltages $V_1$ and $V_2$ are of the same amplitude and of the same phase for simulation of common mode measurement and inverse phase for differential mode measurement. Hence, mixed mode $S$ parameters can be described by formula 6.
Where dm stands for differential and common mode incident and reflected wave denote ports 1 and 2 respectively. Matrix \([S_{dd}]\) represents the differential mode S parameters, \([S_{cc}]\) the common mode S parameters; \([S_{dc}]\) and \([S_{cd}]\) conversion S parameters.

\[
\begin{bmatrix}
    b_{dm1} \\
    b_{dm2} \\
    b_{dm3} \\
    b_{dm4}
\end{bmatrix}
= \begin{bmatrix}
    S_{dd} & S_{dc} \\
    S_{cd} & S_{cc}
\end{bmatrix}
\begin{bmatrix}
    a_{dm1} \\
    a_{dm2} \\
    a_{dm3} \\
    a_{dm4}
\end{bmatrix}
\]

Formula 33 represents an important relationship from which mixed mode S parameters can be determined with a practical measurement system a conventional network analyzer. Practically, it can be measured by the standard single ended S parameters for a balanced circuit, then converting the measured data to mixed mode S parameters. And every the mixed mode measurement in this paper is calculated result from single end measurement of each port.
4 Test Sample

The lines of the test board A and some lines in test board B, with the various lengths, structures and couple factors, were measured in this project, and the results are classified as the transmission, conversion S parameters with SOLT calibration and TRL calibration, difference in group delay and time delay with TDR and TDT method.

4.1 Test Boards

To discuss the material effect, the lines to be tested are on two diverse boards. On test board A, different manufacture factors are presented as standard, offset, and meander. On test board B, lines are assorted as back drilling and un-back drilling.

The specification was schemed in the two tables below.

Table 1: Test sample of test board A \((\varepsilon_{r1}, tg\delta_1)\)

<table>
<thead>
<tr>
<th></th>
<th>Couple factor 1</th>
<th>Couple factor 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Standard line</td>
<td>CF1_line ×</td>
<td>CF2_line ×</td>
</tr>
<tr>
<td>Offset line</td>
<td>CF1_offset_line ×</td>
<td>CF2_offset_line ×</td>
</tr>
<tr>
<td>Meander line</td>
<td>CF1_meander_line ×</td>
<td></td>
</tr>
</tbody>
</table>

where × can be substituted as 1, 2 or 3. It stands as the different lengths, which line 1 is the shortest, line 2 is middle, and line 3 is the longest one.

Table 2: Test board B \((\varepsilon_{r2}, tg\delta_2)\)

<table>
<thead>
<tr>
<th>Transmit line</th>
<th>Receive line</th>
</tr>
</thead>
<tbody>
<tr>
<td>#L_TX_√P_*</td>
<td>#L_RX_√P_*</td>
</tr>
<tr>
<td>#L_TX_√N_*</td>
<td>#L_RX_√N_*</td>
</tr>
</tbody>
</table>

Where # is the number of layer, √ is the number of the pin in the group, and *is the number of the slit on the board. And all the TX lines are with back drilling and RX' are without.

4.2 Lines on the Test boards

First, couple factor 1 and couple factor 2 means two sorts of structure on the same material PCB.
Second, standard lines and offset lines are different in the displacement at the start and end point.

Third, the meander line is another sort of trace.
And at last but not the least, the traces with and without back drilling are showed below.

Figure 41: the meander line structure

A, with back drilling              B, without back drilling

Figure 42: back drilling
5 Results

As mentioned prior, not only material factor effects skew, but the manufactory factors also contribute for this as well, although which doesn’t account as much as the material factor does. And the complications to skew are enumerated as follow both in material aspect and manufactory aspect.

5.1 Result in Material aspect of view - Dielectric constants of the test boards

Based on the modeling in chapter 2, material parameter of the test board A and test board B could be calculated respectively with the single end S parameter of each trace.

![Graph showing dielectric constants of test boards](image)

It is obviously see that the material used on test board A has much lower $\tan\delta$ than test board B has. And the measurement result of mix-mode was showed below.
It schemed that on the test board A which has lower \(tg\), the transmission has 10 to 15dB less insertion loss than the lines on test board B has in the frequency range as 5 to 15GHz. And if we checked the conversion loss,

This figure schemed that the line on test board A has about 10dB conversion loss than the line on test board B has.

Then the suggestion given by the theoretical derive section at the very beginning was proved correct, lower \(tg\) made less dielectric loss, and so that preformed better in the transmission and conversion as the view of mix-mode.
5.2 Measurement result of the Mix-mode

5.2.1 Standard lines and the lines with offset

The standard lines and the offset lines are made of the same length as the test sample section schemed, the only difference is located at the displacement as the figure 47 showed.

Figure 46: standard lines

Figure 47: offset lines

With the couple factor 2, as the previous discussion line 1 was picked up as i.e.,
This picture shows that the standard line presents only about 0.5dB less insertion loss. And refer to the skew which is figured out as conversion loss.

As figure 49 showed that on layer 5, with couple factor 2, the offset line produced about 2dB more conversion loss than the standard line did.

And very similar situation was found in couple factor 1, to contrast with the couple factor 2, the same length was picked up (line1) to discuss.
According to this picture, the similar result with the CF2 lines was obtained that the offset line presented almost the same transmission with the standard line did at 5 to 10GHz, and the standard line showed about 0.5 dB less insertion loss than the offset line on the frequency range from 10 to 18GHz.

And relate to the conversion loss with couple factor 1, it made some difference, that the offset line has about 10dB more conversion loss than the standard line does.
And the situations on others lines, are presented in the appendix attached.
Then the conclusion was made that the offset decreased the transmission and increased the conversion loss because of the discontinuity on the start and end part of the transmission line. And with the different coupling, the offset effects different amount.

### 5.2.2 Couple Factor 1 and Couple Factor 2

Since the line without offset present better transmission and less skew, the following discussion will pick the standard lines as i.e.. Different couple factor means different space and trace width between the pair lines.

![Figure 52: structure of couple factor 1](image)

![Figure 53: structure of couple factor 2](image)

Same with the previous section, line1 was picked out as i.e.

About the transmission,
Figure 54: transmission with different couple factor

It shows that the couple factor 2 line has 0.5dB more insertion loss than the couple factor 1 line.

And refer to the conversion aspect.

Figure 55: difference on the conversion loss

On the figure 55, L5_line1, it is showed about 2dB lower conversion loss in 5GHz to 10GHz with couple factor 2.
To show this effect clearly, the conversion of line2 was showed in figure 56
It is clearer than the result of line1 that there is about 5dB more conversion loss located on the couple factor 2 than it with couple factor 1.

However, this result seems not same with the theoretical conclusion that the strong coupling is better than the lose coupling. For any unbalance such as the length of lines, the width of lines, the placement of the lines, and the material of the board, will be represented on the C, which showed in figure 37. Because the requirement of balance is

\[ C_1 + 2C_M = C_2 + 2C_M \]  \hspace{1cm} (1)

If \( C_M \) is much larger than \( C_1 \) and \( C_2 \), the effect of \( C_1 \) and \( C_2 \) can be ignored, contrarily, the difference of \( C_1 \) and \( C_2 \) will determine the effect of the skew. However, in the measurement result, the loss coupling with smaller \( C_M \) makes better transmission and the conversion. And such phenomenon perhaps was not only account by the couple factor, but also the shape of the line. As the figure 56 schemed that couple factor 1 structure is 2 broad lines with loss coupling, and the couple factor 2 structure was made of 2 thin lines with strong coupling. So not only the coupling factors are different but also the widths of the lines are diverse.
As the figure 57 showed that the material are changing every where, so that the wider line is not as sensitive as the thin line is to the material. In another word, the C1 and C2 in couple factor 1 are much smaller than they are in couple factor 2. So that let $C_M$ compare to them respectively, $C_M$ in couple factor 1 is comparably larger than it is in couple factor 2.

Based on the model built in the preview part, the C of every trace could be calculated with the S parameter of the trace. And the S parameter of one trace was measured when the other trace in the pair was terminated by 50ohm.
And it obviously to see that with couple factor 1 the C is smaller than the capacitance in couple factor 2. However, this calculation is not practical, for when one trace was measured, some signal was coupled to another trace. So such measurement is too rough to be a judgment.

### 5.2.3 Meander line and the standard line

Since couple factor 1 was proved presenting better result in skew part, couple factor 1 was selected as example to discuss the meander line and the standard line.
With the same length, line 1 was picked out as i.e.

![Graph of Insertion Loss](image)

**Figure 60: difference of the insertion loss**

The transmission could be found that they are almost the same. In addition, conversion loss which stands the common response of the device to the differential stimulus was showed that

![Graph of Conversion Loss](image)

**Figure 61: difference of the conversion loss**

It was found that the meander line made about 3dB conversion loss more than the standard line did.
5.2.4 Back drilling or not

Not only the line transmitting signal affected the skew, but some pieces in via holes would be a factor as well. Plated through-hole (PTH) is one of these, which are a means to transport signals into interior layers of a multi layer PCB. PTH vias are used in this way under the pad array of the device package, and in the launch of separable connector interface into boards. The PTH is common to various device packages such as a ball grid array (BGA) and connector types including press fit and surface mount right angle board to board, mezzanine and cable connectors. In the case of a typical backplane system (see Overview Figure 62), the signal passes through at least size vias or PTHs while traveling from driver to receiver.

![Figure 62: Back drilling of PCB](image)

The PTH portion of the signal path becomes more "visible" to the signal as increased frequency content is needed to produce sharp rise and fall times of the digital pulse. And a commonly used way to solve this problem is removing the unused portion of PTH showing following which is referred as back drilling. In this case, RX signal trace is not back drilled, and TX signal trace has back drilling.

Then according to the measurement of 2 pairs of signal trace as RX and TX respectively, some difference could be found as followed. This measurement is with SOLT calibration for TRL calibration needs the CAL KIT. However, in the mix-mode, some problem could be noticed. First, group delay of each trace was checked,
Then it was found clearly on these 2 figures that the 2 pairs of traces are in the same length. And because of they are made in the same material with the same couple factor. It should perform the same transmission and conversion theoretically. In addition, less than 2ns' group delay means the length of these traces are not so long that the term Scd21 can present the conversion loss very well based on the discussion on the previous section. However, in mixed-mode point of view, they are different.
It is clear to see that the trace with back drilling has about 5 dB less insertion loss at the frequency range from 5GHz to 10GHz.

Besides, refer to the conversion diagram, the trace with back drilling has about 10dB less conversion loss than the trace without back drilling dose.
5.3 Contrast to Previous Work

As the method introduction, some previous measurements were also used on this test plane. And as the foregoing analyze, such methods may only be a reference instead of a judgment to these test board, especially at such high frequency as 5 to 15GHz. Since the results in the former section that some manufacture problem such as offset, meander, and distinct coupling factor, even some path in via hole, could make skew were obtained, the contrast to the previous work in the following part seems worth.

First, start by the test board A.

<table>
<thead>
<tr>
<th>Name of line</th>
<th>TDR (ps)</th>
<th>TDT (ps)</th>
</tr>
</thead>
<tbody>
<tr>
<td>CF1_line1</td>
<td>4.4</td>
<td>7.8</td>
</tr>
<tr>
<td>CF1_line2</td>
<td>2.5</td>
<td>8.6</td>
</tr>
<tr>
<td>CF1_line3</td>
<td>7</td>
<td>5</td>
</tr>
<tr>
<td>CF2_line1</td>
<td>2</td>
<td>1.1</td>
</tr>
<tr>
<td>CF2_line2</td>
<td>3.5</td>
<td>8.9</td>
</tr>
<tr>
<td>CF2_line3</td>
<td>2</td>
<td>1.43</td>
</tr>
<tr>
<td>CF2_offset_line1</td>
<td>1.6</td>
<td>9</td>
</tr>
<tr>
<td>CF2_offset_line2</td>
<td>3</td>
<td>5.5</td>
</tr>
<tr>
<td>CF2_offset_line3</td>
<td>6</td>
<td>6.9</td>
</tr>
<tr>
<td>CF1_line1</td>
<td>1.6</td>
<td>7</td>
</tr>
<tr>
<td>CF1_line2</td>
<td>5</td>
<td>1.39</td>
</tr>
<tr>
<td>CF1_line3</td>
<td>1</td>
<td>1.62</td>
</tr>
<tr>
<td>CF1_offset_line1</td>
<td>5.6</td>
<td>3.7</td>
</tr>
<tr>
<td>CF1_offset_line2</td>
<td>0.5</td>
<td>9</td>
</tr>
<tr>
<td>CF1_offset_line3</td>
<td>4</td>
<td>1.78</td>
</tr>
</tbody>
</table>

If we compare the data from the two methods, they are not in the same trend, but they should give some similar result. And if we compare the data of each group of lines, we cannot find any rule to conclude them. That means it's hard to judge the skew in time domain. As the disadvantages of TDS we discussed in the previous section, Then relate to the test board B.

<table>
<thead>
<tr>
<th>Name of line</th>
<th>TDR (ps)</th>
<th>TDT(ps)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4L_RX0PN_2</td>
<td>9.6</td>
<td>2.8</td>
</tr>
<tr>
<td>4L_RX2PN_2</td>
<td>-12.4</td>
<td>10.4</td>
</tr>
<tr>
<td>4L_RX3PN_2</td>
<td>5.4</td>
<td>12.8</td>
</tr>
<tr>
<td>4L_TX0PN_2</td>
<td>2</td>
<td>11.6</td>
</tr>
<tr>
<td>4L_TX2PN_2</td>
<td>-0.4</td>
<td>11</td>
</tr>
<tr>
<td>4L_TX3PN_2</td>
<td>3.6</td>
<td>12.8</td>
</tr>
</tbody>
</table>
Table 2 schemed 2 group of line, in which one of them is with back drilling, and the other is without. And the lines in the same group are made in the same length. So the data in one group should be some more or less value. And further more, according to the new method in this paper, the effect of back drilling has been showed in the previous section, and with the previous method, as the table 2 presented, not any clear rule could be concluded.

At last, group delay in frequency domain was showed below.

Even not any clear result can be seen. And in this pair of line, actually relied some skew (all the measurement result could be checked in the appendix).

Based on such contrast, the problem could be found out that modern digital designs and verifications require knowledge that has formerly not been needed for a data bit rate of below than 100Mbit per second. So at the higher frequency range as 5 to 15GHz, most references on the necessary subjects are too abstract to be immediately applicable to the skew.
5.4 Simulation work

Advanced Design System (ADS) is an electronic design automation software system produced by Agilent EEsOf EDA [1], a unit of Agilent Technologies. It provides an integrated design environment to designers of RF electronic products such as mobile phones[2], pagers, wireless networks, satellite communications, radar systems, and high-speed serial links[3].

Agilent ADS supports every step of the design process -- schematic capture, layout, frequency-domain and time-domain circuit simulation, and electromagnetic field simulation -- allowing the engineer to fully characterize and optimize an RF design without changing tools.

Agilent EEsof has donated copies of the ADS software to the electrical engineering departments at many universities, and a large percentage of new graduates are experienced in its use. As a result, the system has found wide acceptance in industry.

The disadvantage, however, is that the material of the PCB in ADS is uniform which could not realize on the hardware, so all the standard line in simulation present only conductor loss in transmission, and no loss in conversion.

Here some simulation work was done by ADS for the standard line and the offset line with different couple factors. All the simulation was based on the test board A, with the length of line1.

First with couple factor 1, the simulation output was schemed that

![Figure 68: Simulation of differential mode transmission of CF1](image)

It was showed that the offset made about 1 to 5dB more insertion loss than the standard line did in the frequency since 5GHz to 20GHz because of the discontinuity of the impedance on the start and end part of the transmission line. And relate to the conversion part.
As the material in simulation is uniform, for a perfect match coupled line, there is not any conversion loss on the standard line. And then with couple factor 2, the result was similar.

There is about 1 to 6dB more insertion loss with the offset line in the frequency range as 5 to 20GHz. And relate to the conversion
Figure 71: Simulation of differential mode conversion of CF2

Almost same result to the CF1 group did.

At last, contrast of CF1 and CF2 standard line was showed

Figure 72: Simulation of differential mode transmission of CF1 and CF2

It was showed not any difference. However, the measurement schemed that couple factor 1 preformed a better result on transmission. This is because the material in ADS simulation is uniform, the material changing effect to the transmission line could not be found in this simulation. For this reason, if the relation between the width of line and the asymmetry of material will be study in the future work, a new board with different width line is required.
6 Conclusion

According to the previous discussion, some conclusion statements are enumerated as follow.

1. Using Mix-mode is able to check the skew in frequency domain, especially in high frequency range.

2. Scd21 is the best choice to find out the skew in high frequency range.

3. TRL calibration is able to eliminate the connector effect, so that the skew made by discontinuity could be checked.

4. In material aspect, lower tangent made less dielectric loss and produced smaller conversion loss.

5. Not only the couple factor related to the unbalance, but also the width of line.

A comment should be lodged that the result 4 and 5 above have been misunderstanding for many years as the connector jam instead of the common mode brought by the manufacturer [13]. However, this conclusion needs to be confirmed by further laboratory work. Although more test board is required, however, such factor is different from the theoretical knowledge, so it is worth to do in the future work.
7 Reference

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