[Technical Paper]

Evaluation of Imbalance Component and EM Radiation Generated by an Asymmetrical Differential-Paired Lines Structure

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Abstract

Recently, differential-signaling (DS) techniques such as low-voltage differential-signaling (LVDS) have been widely used in digital electronics devices in order to suppress electromagnetic interference (EMI). But in practical terms, a complete topologically and structurally symmetrical differential line is impossible. In this paper, we newly attempt to quantify the imbalance component and electromagnetic (EM) radiation when the structure and topology change from a symmetrical to an asymmetrical differential paired lines. Four different differential-paired lines structures are prepared for the comparison: PCB1 is a basic symmetrical structure taken as an “ideally balanced” case, PCB2 is an asymmetrical structure due to differences in bend and length, PCB3 is a symmetrical length structure with a bend region, and PCB4 is an asymmetrical topology with equi-distance and bend routing. Firstly, the differential voltages and mixed-mode scattering parameters are selected as a measure. The conversion parameter from differential-mode (balance component) to common-mode (imbalance component), S_{cd21} , is dramatically increased by the difference of the length. Even if the differential paired lines have a bend-region, equi-distance routing can suppress S_{cd21} . Secondly, spatial distributions of near magnetic fields are measured at certain resonant and out of resonant frequencies of S_{cd21} . Although the differential paired lines are excited by the differential-mode, propagated magnetic field component at the end terminal of the differential paired lines at the resonant frequencies of S_{cd21} could be changed to the common-mode. Thirdly, the far-electric fields at 3 m are measured and calculated. Even if equi-distance routing is suitable for the improvement of signal integrity (SI) issues, it is not enough for the suppression of the far-field potential radiation. It is clear that S_{cd21} is one evaluator but it is not sufficient for predicting the EM radiation completely. The facts shown in this study suggest the basic characteristics of EM radiation from practical differential paired lines with asymmetrical structure and some significant problems in the design of a meander delay line for high-speed clock distribution.

Keywords: Electromagnetic Interference, Electromagnetic Radiation, Common-Mode, Differential-Mode, Differential-Signaling

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1. Introduction

In recent years, the clock frequency and packaging density of electronic devices have continued to increase, resulting in more potential noise interference and radiation issues. All electronic devices must satisfy the requirements of electromagnetic compatibility (EMC) for low electromagnetic interference (EMI).[1] Hence, effective methods for predicting and suppressing EMI as well as maintaining signal integrity (SI) over a broad band are required.

Differential-signaling (DS) techniques such as low-voltage differential-signaling (LVDS) are widely used in digital

electronic devices in order to suppress EMI.[2–9] Improvements of SI performance such as cross-talk noise and eye-diagrams have been reported. In addition, the suppression effect of the differential-transmission mode on electromagnetic (EM) radiation has been demonstrated. [6] So far, the authors have discussed the characteristics of the EM radiation from a printed circuit board (PCB) driven by LVDS, using three-EMI antenna mode, which can identify and explain the frequency response of the EMI.[10, 11] For actual differential signaling, the ideal balance or symmetrical structure and topology cannot be established; hence, an imbalance component should be

excited in practical high-density packaging systems.[3, 10, 12] Many papers have demonstrated that imbalance components in differential signals deteriorate EMI and SI performance. For mitigation of EMI problems underlying differential paired lines, the identification of the primary factors which generate the imbalance component should be an important issue in order to quantify the relationship between the EMI and the configurations at the design stage for practical applications.

Imbalance components between voltages traveling the differential paired lines are generated by the non-ideal symmetrical structure or topology, and discontinuity of the differential-mode (DM) impedance of the differential paired lines, and the skew and distortion of the output

waveform at the signal driver. This paper focuses on the imbalance components generated by the asymmetrical structure and topology, which depend on the PCB layout, and which are important issues in the design stage. In order to provide the basic considerations for EM radiation from practical differential paired lines with typical asymmetrical structure or topology, we newly attempt to quantify the imbalance component and EM radiation from an asymmetrical differential paired line driven by LVDS. Firstly, the PCB geometries used in the study are described in Section 2. In Section 3, the differential voltages, mixed-mode scattering parameters, and magnetic near fields are measured and calculated. Then, the imbalance component on the differential paired lines is quanti-

Fig. 1 Geometry of the PCB under study (in mm).

fied and discussed. In Section 4, the far-electric fields at 3 m are discussed by experiment and calculation. Specifically, the focus is on the correlation between the imbalance component and the EM radiation. Section 5 is conclusions.

2. PCB Geometry Under Study

The differential-paired lines with four different layouts were prepared for the discussion as typical structures and topologies. The geometries of the PCBs under study are illustrated in Fig. 1. 1) is a basic symmetrical structure as the “ideally balanced” case, called PCB1; 2) is an asymmetrical structure due to differences in bend and length, called PCB2; 3) is a symmetrical length structure with a bend region, called PCB3; and 4) is an asymmetrical topology with equi-distance and bend routing.

The PCB has two layers, with the upper layer for the signal trace and the lower layer for the reference (ground) plane. The size of the microstrip line structure used for the test model is $l = 137$ mm (length), $w = 100$ mm (width), and

$h = 1.53$ mm (thickness) for the dielectric substrate with a permittivity of $\epsilon_r = 4.5$. Two traces with $l = 100$ mm, width w_t , and distance s are located on the dielectric substrate. As the focuses in this paper are on the imbalance component generated by asymmetrical structures and topologies, the relatively wide separation $s = 1.0$ mm is selected so that an imbalanced component due to the asymmetrical structure is the dominant fact of the EM radiation compared with that due to the waveform distortion of the output of the LVDS driver. PCB1 has a symmetrical differential paired lines with the differential mode impedance $Z_{DM} = 100 \Omega$. Table 1 shows the characteristic impedance of the differential paired lines in each region.

Table 1 Characteristics impedance of differential paired line.

s [mm]	w_t [mm]	Z_{DM} [Ω]	Z_{CM} [Ω]
1.0	1.9	100	36.3
26.0		125	31.2
51.0		125	31.2

(a) PCB1

(b) PCB2

(c) PCB3

(d) PCB4

Fig. 2 Waveform of voltage on differential paired lines measured at the end of the lines. V_1 and V_2 : the line voltages, $V_{DS} = V_1 - V_2$: differential-mode voltage, $V_{CM} = (V_1 + V_2)/2$: common-mode voltage.

3. Imbalanced Component on Differential Paired Lines

3.1 Model Evaluation of Voltage Waveform

Figure 2 shows the voltage waveforms on differential paired lines measured at the ends of the lines. The differential paired lines are driven by an LVDS driver (NS DS90LV047A), and a crystal oscillator with an amplitude of 3.3 V and a frequency of 25.0 MHz. The differential paired lines are terminated with $100\ \Omega$ to match the impedances at the end of the lines, port 2. The voltages traveling Line 1 and Line 2 are asymmetrical waveforms due to the differences in the phase-shifts and amplitudes. Hence, the imbalanced, common-mode (CM), current arises. The fluctuation of the CM voltage in PCB2 is 122 mV and is relatively large compared with the other cases: 76 mV for PCB1, 71 mV for PCB3, and 82 mV for PCB4. An appropriate estimate of the imbalance due to the asymmetrical waveform is the key to the prediction of the total SI/EMI behavior of the PCB driven by LVDS.

Figure 3 shows the simulated eye-diagram for the differential-voltage, V_{DS} , for the 2.5 GHz signal. Table 2 shows the simulated maximum eye width (MEW) as an evalua-

Table 2 Evaluation of eye-diagram.

PCB	Maximum eye width (MEW) [ps]
PCB1	396
PCB2	210
PCB3	396
PCB4	374

tion of the eye-diagram. For the ideally balanced case, PCB1, the MEW is 400 ps. For PCB2, the MEW is 210 ps and is dramatically deteriorated by the different length. The MEW for PCB4 is 374 ps. This indicates that equi-distance routing is suitable for the improvement of signal integrity issues such as the eye-diagrams of high-speed digital circuits.

3.2 Evaluation of Mixed-Mode S-parameter

The frequency responses of mixed-mode scattering parameters are shown in Figs. 4 and 5. The solid and dashed lines show the measured results from the network analyzer with 4 ports, and the FDTD calculated results, respectively. Figure 4 shows the frequency response of $|S_{dd21}|$, which is defined as the transmission coefficient of

Fig. 3 Simulated eye-diagram of differential-voltage V_{DS} for 2.5 GHz signal.

Fig. 4 Frequency response of $|S_{dd21}|$. Fig. 5 Frequency response of $|S_{cd21}|$.

the differential-mode. The $|S_{dd21}|$ of PCB2 is deteriorated compared with that of the other three models.

Figure 5 shows the frequency response of $|S_{cd21}|$, which is defined as the conversion from differential-mode (balance component) to common-mode (imbalance component). $|S_{cd21}|$ is dramatically increased by the different lengths of Line 1 and Line 2. Additionally, even if the structure is asymmetrical differential paired lines, equi-distance routing can suppress the $|S_{cd21}|$. The resonant and anti-resonant frequencies of PCB2 correspond to the phase differences due to the different lengths of Lines 1 and 2.

As the proposed model for predicting the frequency response of $|S_{cd21}|$ in PCB2 is based on the transmission line theory, the differential-mode currents at the ends of the differential paired lines can be calculated by the following:

$$I_1 = I_{a1} e^{-(\alpha + j\beta)l_1}, \quad (1)$$

$$I_2 = -I_{a2} e^{-(\alpha + j\beta)l_2}, \quad (2)$$

where I_{a1} and I_{a2} are the magnitude of the differential-mode currents, $l_1 = 100$ mm, and $l_2 = 150$ mm are the lengths of Signal Lines 1 and 2, respectively, and β is the phase constant which is calculated by an assumption of single-ended transmission.

The imbalance component current at the end of the differential paired lines can be calculated by

$$I_{imb} = I_{a1} e^{-(\alpha + j\beta)l_1} - I_{a2} e^{-(\alpha + j\beta)l_2}. \quad (3)$$

For a lossless transmission line ($\alpha = 0$) and ideally balanced excitation ($I_{a1} = I_{a2} = I_a$) case, it is derived simply as a function of the phase difference between Lines 1 and 2:

$$I_{imb} = I_a e^{-j\beta l_1} - I_a e^{-j\beta l_2}. \quad (4)$$

Figure 6(a) shows the phase frequency response of each DM current at the end of the differential paired lines, and (b) shows the frequency response of the phase difference of the currents between Lines 1 and 2, and the magnitude of the imbalance component current. When the phase difference is 0 degrees (in-phase), the imbalance current becomes a maximum. On the other hand, the imbalance current is null when the phase difference is 180 degrees (reversed phase). The resonant and anti-resonant frequencies obtained by measurement and calculation are shown in Table 3. The measured and predicted frequencies are in good agreement, within approximately 10%. The proposed model can explain the frequency responses of the imbalance component driven by differential signaling and also

(a) Frequency response of phase of each differential-mode current at the end of the differential paired line

(b) Frequency response of the phase difference of currents between lines 1 and 2, and magnitude of imbalance component current.

Fig. 6 Predicted results of frequency response of imbalanced component in the PCB2 case.

Table 3 Comparison between measured and predicted results.

Frequency	Measurement [GHz]	Prediction [GHz]	Error [%]
First resonant	1.85	1.69	8.6
First anti-resonant	3.78	3.33	11
Second resonant	5.40	5.03	6.9
Second anti-resonant	7.55	6.67	11

predict the resonant and anti-resonant frequencies.

3.3 Evaluation of Magnetic Near Field

To understand the details of the difference in the frequency characteristics of the imbalanced component S_{cd21} , spatial distributions of the magnetic near field on the dif-

ferential paired lines are one specific measures. The differential paired lines are driven by an LVDS driver (NS DS90LV047A), and a crystal oscillator with an amplitude of 3.3 V and a frequency of 25.0 MHz. The differential paired lines are terminated with 100 Ω loads to match imped-

Fig. 7 Magnetic field distribution at 2.0 mm above the differential paired lines (25 MHz).

ances at the end of the lines, Port 2. The magnetic near fields, H_x and H_y , were measured using a loop-coil type magnetic near-field probe with a $250\ \mu\text{m}$ spatial resolution and were calculated as $|H|$.

Figures 7 and 8 show the magnetic near field $|H|$ distribution at about 2.0 mm above the surface of the differential paired lines at 25 MHz (fundamental frequency), and 1850 MHz (resonant frequency of S_{cd21} for PCB2), respectively. At 25 MHz (Fig. 7), the magnetic near field is mainly distributed on each trace, Line 1 and Line 2, because of differential signaling. Also, the magnetic field strength is almost constant from Port 1 to Port 2. At the higher frequency, 1850 MHz (Fig. 8), the magnetic field strength is

no longer constant. Although Logical Port 1 of the differential paired lines are excited by the differential-mode, the field component at Logical Port 2 of the differential paired lines at the resonant frequency of S_{cd21} is common-mode. In PCB4, there is a region of even-mode propagation at the center position of the differential paired line ($x = 0$). Such even-mode propagation generates a CM current and may result in significant radiation as a CM dipole.

4. Results and Discussion on Radiated Emission

Experimental and full-wave analyses based on the FDTD method are performed to discuss the correlation between the imbalance component and EM Radiation from the

Fig. 8 Magnetic field distribution at 2.0 mm above the differential paired lines (1850 MHz).

Fig. 9 Experimental setup for far-electric field measurement.

PCB. The differential paired lines are driven by an LVDS driver (NS DS90LV047A), and a crystal oscillator with a 3.3 V amplitude and 25.0 MHz oscillation frequency. The crystal oscillator, LVDS driver, and battery are installed on the reverse side, and are covered with a shield box (thin-sheet copper). The differential paired lines are terminated with a 100 Ω SMT resistor.

The far-electric field radiated from the PCB was measured in an anechoic chamber with a Bi-Log antenna (30–1000 MHz), as shown in Fig. 9. The coordinate system is also notated, where ϕ is the counterclockwise angle in the x - y plane measured from the positive x axis, and θ is the angle from the positive z axis. The PCB was placed on a wooden table. The distance between the PCB and the receiving antenna was 3 m. The RBW and VBW of the spectrum analyzer are 100 kHz. The horizontal component, E_ϕ at $(\phi, \theta) = (90^\circ, 90^\circ)$: E_x , and the vertical component, E_θ at $(\phi, \theta) = (90^\circ, 90^\circ)$: E_z , are discussed.

The measured far-electric fields shown in Figs. 10 (a) and (b) are the horizontal component, E_x , and the vertical component, E_z , respectively. The measured results which are smaller than the noise floor level are omitted in Fig. 10.

The horizontal component, E_x , is larger than the vertical component, E_z . The radiation in PCB2 is the largest in all four conditions. The radiations in PCB3 and PCB4 are large compared with the radiation in PCB1. This fact does not correspond to the feature of the imbalance component evaluated from the magnetic field. Although equi-distance routing is suitable for improvement of SI issues, it is not effective in suppressing the EMI. More importantly, the result indicates that the measurement of S_{cd21} is not a sufficient parameter for predicting the EM radiation. These facts are significant problems in the design of a meander

 (a) Horizontal Component E_x

 (b) Vertical component E_z

Fig. 10 Measured frequency responses of the far-electric field.

delay line for high-speed clock distribution and the basic characteristics of EM radiation from practical differential paired lines with an asymmetrical structure.

Figure 11 shows the angle pattern of the far-electric field measured at 375 MHz, which is the frequency at the highest radiation. The horizontal component, $E_\phi(H)$, in PCB1 results from CM radiation driven as a half-wavelength dipole antenna. Therefore, the angle pattern of E_ϕ seems

Fig. 11 Angle pattern of the far-electric field measured at 375 MHz. $\theta = 90$ degrees and $\phi = 0$ to 360 degrees. $E_\phi(H)$: horizontal component, $E_\theta(V)$: vertical component.

like a bidirectional-sinusoidal pattern, E_ϕ has directivity at $\phi = 90$ and 270 degrees, and $\phi = 0$ and 180 degrees are null. The vertical component, $E_\theta(V)$, in PCB1 is due to the DM loop antenna with directivity at $\phi = 0$ and 180 degrees. The angle patterns of asymmetric cases are more complex. It is difficult to identify the dominant radiation factor of the other layouts (PCB2, 3 and 4) by using the simple dipole and loop type antennas.

The FDTD method is used for the calculation of the far-electric fields. The FDTD model provides results for the “ideal balanced differential-signaling excitation, ($V_1 = -V_2$),” which can illuminate the effects of non-ideal asymmetrical structures and the discontinuity of DM impedance of differential paired lines. The condition of FDTD modeling was determined from our previous studies.[11] By using the near-to-far-field transformation, the far-electric field was obtained from the surface electric and mag-

netic current distributions on the closed surface surrounding the FDTD model. Figure 12 shows the frequency response of the far-electric field, obtained by FDTD modeling. To discuss the frequency response of the far-electric field from the point of view of radiation efficiency, the result is normalized to the case where the frequency response of the input voltage to the trace is a constant 0 dBm. The horizontal component is the CM radiation resulting from the non-zero impedance due to the finite ground plane. The vertical component results from the uncanceled component at the observation point due to the separation between the two lines. As the frequency becomes higher, EM radiation increases. A feature of the frequency response is similar to that of the measured results: the equi-distance routing is not effective in suppressing the EMI.

Three factors could be possible reasons why equi-dis-

(a) Horizontal Component E_x (b) Vertical component E_z

Fig. 12 Calculated frequency responses of the far-electric field in the case of “ideal balanced differential-signaling excitation ($V_1 = -V_2$).”

tance does not suppress EMI.

- 1) The differential-paired lines are routed in the edge of the PCB in the model layout used. So, the CM current due to the finite impedance of the ground plane increases.
- 2) The distance between the differential-paired lines, the space between Line 1 and Line 2, becomes wider. Then, the cancellation effect for EM radiation at the observation point deteriorates dramatically. The relationship between distance, s , and theoretical improvement, α , with differential signaling, where α is defined as the ratio of the far-electric field radiated from differential pair lines to that from a single-end microstrip line:

$$\alpha \equiv \frac{E_{DS}}{E_{single-end}} = 1 - \frac{r}{r + s_c} e^{-j \frac{2\pi s_c}{c_0}}, \quad (5)$$

where s_c is the distance between the center of two traces, $s_c = s + w$, c_0 is the speed of light, and r is the distance from Line 1 to the observation point at the far-electric field measurement.[6] As the distance, s , between the differential paired lines increases, α deteriorates.

- 3) There is a spatial region of even-mode current propagation due to the asymmetric layout. This results in a significant CM current and radiation as a CM dipole.

Although the results do not allow us to distinguish the particular radiation component from each of these factors, these consequences indicate that the equi-distance routing does not work as an effective factor for suppressing EMI. Predicting the dominant component at a certain frequency band is a future subject for mitigating EMI problems in high-speed electronic designs.

5. Conclusion

In order to provide basic considerations for the realization of methods for predicting and suppressing the EM radiation from PCBs driven by LVDS, the imbalance component and EM radiation from such PCBs were studied experimentally and with numerical simulations. This paper reported the basic characteristics of EM radiation from practical differential paired lines with a typical asymmetrical structure. Although equi-distance routing is suitable for the improvement of SI issues, it is not sufficient in suppressing the EMI. More importantly, the results indicate that the measurement of S_{cd21} is not the only evaluator for predicting the EM radiation. As the discussions here are limited to differential paired lines with wide separation, the acquired knowledge can be applied to narrow separation (more practical) cases.

Future work includes studies of engineering implications, development of easier interpretations for coupling and CM-generation mechanisms, and the identification of the dominant radiation factor.

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