Modeling and Analysis of Vertical Noise Coupling in TSV-Based 3D Mixed-Signal Integration

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Abstract

This paper reports on modeling and analysis of vertical noise coupling between adjacent chips in TSVs-based 3D mixed-signal integration. The coupling between clock channel in digital IC and channel routing wires (CRWs) in analog IC is studied in both frequency domain and time domain. According to physical architecture, a broadband equivalent circuit model of vertical noise coupling is proposed. The comparison of S parameters from full-wave electromagnetic (EM) simulation and circuit model shows a good agreement. The coupling mechanism is discussed from transfer impedance between clock channel and CRWs. The impacts of design variables (such as geometry dimension, 3D stacking process, layout floorplan) on noise coupling are investigated. Four approaches are proposed to reduce the vertical noise coupling and compared with the baseline model. Finally, the time domain analysis shows that the rising/falling time determines peak-to-peak noise voltage and the far-end exhibits larger noise than the near-end.

Keywords: Channel routing wire; Clock channel; Equivalent circuit model; Mixed-signal; Through silicon vias (TSVs); Vertical noise coupling.

1. Introduction

Motivated by the requirements of high speed, high bandwidth and high power efficient electronic systems in modern society, three-dimensional (3D) integration technology based on vertical interconnects that penetrate through silicon strata was developed [1]-[4]. It is promising to be applied into fields like memory [5]-[7], image sensors[8]-[10] and microprocessors[11][12] for it involves lower interconnect latency, lower power consumption, smaller form factor and easier for heterogeneous integration between different technologies.

In 3D mixed-signal integration systems, analog circuits, digital circuits and sensors are usually vertically stacked. Kiyoyama et al [13] has proposed a 3D stacked image processing system, in which, a parallel analog digital convertor (ADC) with hierarchical correlated double sampling technique is vertically stacked with image processing circuit by TSVs. Liu et al [14] reports the design of a TSV based three-layer stacked 12-bit successive approximation register (SAR) ADC, where the analog parts of ADC are stacked on top of digital parts. Since analog circuits such as readout integrated circuits (ROIC) and ADC are sensitive to the switching noise from digital circuits, it is essential to evaluate the characteristics of vertical coupling behavior between analog circuits and digital circuits in mixed-signal 3D integration systems, especially for state-of-the-art 3D integration scheme, in which, silicon substrates are thinned down to about 50 μm or less and vertical noise coupling between heterogeneous chips becomes more severe. All in all, by applying TSV technology into 3D mixed-signal integration systems, relatively shorter signal path and reduced

form factor can be obtained, but on the other hand, it raises concerns about vertical noise coupling from digital to analog because of the thin chip substrate and high switching frequency. Therefore, the demands of investigation about vertical noise coupling in TSV-based 3D mixed-signal integration will be urgent.

Previous works have shown that vertical noise coupling between adjacent chips are significant to performance degradation in 3D devices [15][16]. However, these studies have focused on the analysis of vertical crosstalk in 3D system in which the connections between adjacent chips are realized through wire bonding. In contrast, this work provides the detailed analysis of the vertical noise coupling in TSV-based 3D mixed-signal integration. By applying finite element methods (FEM), EM field is solved to capture parasitics and couplings. The corresponding broadband equivalent circuit model is proposed and compared with EM solution up to 40GHz. Moreover, the coupling mechanism and impacts of design variables are analyzed in detail, based on which four methods are proposed to mitigate the noise coupling. Furthermore, the transient noise transfer characteristics such as ringing effect and peak-to-peak noise voltage at both near-end and far-end are also investigated.

2. Modeling

2.1. TSV-Based 3D Architecture for Mixed-Signal Integration

In synchronous digital systems, clock signal is used as time reference for data processing, typically passes through the longest distance and works at the high speed [17]. The clock distribution is to deliver global clock across the chip to reach all of the clocked elements [18]. In analog or mixed-signal layout design, a technique that routing the wires alongside the individual circuit blocks is known as channel routing [19]. Fig.1 shows a baseline 3D structure employed to model the vertical noise coupling between the clock channel and channel routing wires (CRWs). The baseline architecture is constituted of a top analog chip and a bottom digital chip bonded together with benzocyclobutene (BCB). The clock channel is realized using the top metal layer (M9) embedded in inter-metal dielectric layer 2 (IMD2) of the bottom tier while the CRWs (labeled as ‘a’ and ‘b’) are made of M3 embedded in IMD1 of the top tier. The ground grids made of M8 and M9 in each tier are connected with six TSVs. The geometry and material parameters are listed in Table I, which are based on TSMC 90 nm CMOS technology. Clock_M9 and CRW_M3 represent the clock channel and CRWs, respectively. Ground_M9 and Ground_M8 connected by Via_M8_M9 form the ground grids. Substrate (T) and substrate (B) stand for the top and bottom substrate, respectively. The area of substrate ($600\mu\text{m} \times 600\mu\text{m}$) is chosen by considering the typical size of building block in mixed-signal circuit design. And the initial width, length, and space of metal wires are determined from common practice in mixed-signal circuit design, while the impact of geometry dimension on noise coupling is analyzed in section 3. The thicknesses of the top substrate, bottom substrate, and bonding material as well as the geometry of TSVs are chosen based on the realistic fabrication process of 3D integration.

Fig. 1. (a) 3D view of baseline model with two stacked chips.

(b) Sideview of baseline model for simulation (not in scale).

TABLE I
DIMENSION AND MATERIAL PARAMETER OF BASELINE MODEL

Name	Width (μm)	Length (μm)	Thickness (μm)	Space (μm)	Relative permittivity	Conductivity (S/m)
Clock_M9	10	600	3.4	-	1	5.88×10^7
CRW_M3	5	600	0.31	6	1	4.48×10^7
Ground_M9	10	600	3.4	190	1	5.88×10^7
Ground_M8	10	600	0.85	190	1	5.35×10^7
Via_M8_M9	10	10	0.75	190	1	1.49×10^7
Substrate(T)	600	600	40	-	11.9	10
Substrate(B)	600	600	100	-	11.9	10
IMD1	600	600	4.5	-	2.9	0
IMD2	600	600	7.5	-	4.2	0
Bonding material	600	600	5	-	2.6	0
TSV	5(Diameter)	53.6	-	195	1	5.88×10^7
Insulator	-	53.6	0.3	195	3.9	0

2.2. EM Solution

ANSYS's HFSS is employed as a 3D EM field solver to capture parasitics and simulate vertical noise coupling between the clock channel and CRWs in the baseline model. All ports are terminated with 50Ω and the ports allocation are shown in Fig. 2(a). Due to the symmetry of CRWs 'a' and 'b' with respect to the clock channel, the near-end crosstalk (NEXT) S_{31} and far-end crosstalk (FEXT) S_{41} are equal to S_{51} and S_{61} , respectively. As shown in Fig. 2(b), the fluctuation of S_{11} at high frequency indicates the self-resonance occurs between the parasitic capacitance and the self-inductance of clock channel. Besides, NEXT and FEXT increase with frequency until approaching the self-resonant frequency, and the crosstalk (S_{31} and S_{41}) is larger than -30 dB above 3 GHz, namely, a 1-V signal transmitted through the clock channel can generate more than 31 mV noise at the CRWs, which may overwhelm the small analog signal and severely degrade the performance of the analog chip.

Fig. 2. (a) Ports allocation of the baseline model. (b) Simulated reflection coefficient (S_{11}), NEXT (S_{31}) and FEXT (S_{41}) from 0.1GHz to 100GHz.

2.3. Coupling Mechanism

Capacitive coupling through parasitic capacitance and inductive coupling through mutual inductance are the two typical coupling paths between clock channel and CRWs. In order to understand the dominant coupling mechanism of the baseline model, transfer impedance is extracted from the EM solution, which represents the coupled noise voltage in the CRW induced by a unit current injected into the clock channel. As shown in Fig. 3(a), the transfer impedance increases linearly with frequency and exhibits a pure inductive characteristics from 0.1 GHz to 15 GHz, where the capacitive coupling can be ignored. Fig. 3(b) depicts the mutual inductance derived from transfer impedance, which is constant with 0.16 nH up to 15 GHz. This indicates that the magnetic fields generated by currents flowing through the clock channel and CRWs strongly couple with each other, which makes the inductive coupling become the dominant path. The capacitive coupling starts to play a role at frequency above 15 GHz because the transfer impedance increases faster than linear with frequency, in other words, the equivalent mutual inductance increases with frequency, as can be seen from Fig. 3(b). The capacitive coupling dominates beyond 40 GHz where the transfer impedance decreases with frequency and the equivalent mutual inductance becomes

negative.

Fig. 3. (a) Transfer impedance between clock channel and CRW and (b) Simulated mutual inductance between Port 3 and Port 1 of the baseline model.

2.4. Equivalent Circuit Model

Although the S parameter files obtained from the EM simulation can be directly used as a blackbox in the commercial IC design tools such as Cadence, the complex nature of S parameters due to various parasitic effects makes the time-domain and large-signal simulation like periodic steady state (PSS) or periodic noise (PNoise) very time consuming and even prone to potential convergence problems [20]. Therefore, it is highly desired to build a physics-based lumped element model that can characterize the vertical noise coupling behavior and help in simulating the 3D mixed-signal integrated circuits. From a practical point of view, the equivalent circuit model also needs to be applicable over wide bandwidth, because the sharp rising and falling edge of the clock signal in digital chip contains high frequency components. For instance, to capture the vertical noise coupling generated by a 3 GHz square wave clock signal with 33ps raising time (10% of signal period), the model should at least cover the 7th-order or 9th-order harmonic of the fundamental frequency, which means the model should be valid from DC to nearly 30 GHz.

Fig. 4(a) shows a basic π model that are commonly used for a metal trace routing above the lossy substrate. In this work, two methods are employed to further broaden the bandwidth, first, $2-\pi$ equivalent circuit is adopted which can better described the distributed feature of the long clock channel and CRWs, second, the series R-L in π model is replaced by a ladder network consisting of L , L_f , R_f and R_m in order to capture frequency dependent loss from the metal due to skin effect [20]. As illustrated in Fig. 4(b), the clock channel, CRW ‘a’ and ‘b’ are modeled using three $2-\pi$ networks separately. The inductive coupling through mutual inductance and capacitive coupling through top substrate among clock channel, CRW ‘a’ and ‘b’ are then added to the $2-\pi$ network, while the common ground reference for the whole structure is setting at the top and bottom ground grid connected by multiple TSVs. The physical meaning of each component is explained as follows.

The self-inductances of each CRW and clock channel are represented by series inductance L_{signal} and L_{clock} , respectively, and M_{CS} stands for the mutual inductance between them. C_{ss} and M_{ss} represent the capacitance and mutual inductance between the two parallel CRWs. The capacitance and resistance of top silicon substrate are modeled by C_{sub} and R_{sub} in parallel, which are determined from the dielectric relaxation time of the substrate [21]. The oxide capacitance between the CRWs and the silicon substrate is modeled by C_{IMD1} , and C_{bond} is the capacitance between clock channel and top substrate due to the bonding layer. C_{signal} and C_{clock} model the capacitance from CRWs and clock channel to ground grids, respectively. The values of multiple parameters in the equivalent circuit can be obtained through two steps, first, a set of estimated values are derived directly from analytical calculation [21], which are used as the initial values for parameter fitting and optimization during the second step. Since the variation of the parameters are confined within a limited range around their calculated values based on the physical dimensions and material property, it prevents the iteration from converging to a local minimum or resulting in physically meaningless values. The corresponding parameters of each component are listed in Table II. In order to verify the accuracy of proposed equivalent model, the simulated S parameters from the lumped element model and 3D full-wave EM solver (HFSS) were compared in Fig. 5(a)-(d) over wide bandwidth (i.e. 0.1 GHz- 40 GHz). It can be seen that, good agreement is achieved for both reflection coefficient (S_{11} and S_{33}) and coupling coefficient (S_{31} and S_{41}), which proves the broadband nature and accuracy of the proposed model. And the obtained model can be utilized in circuit design software for almost any kinds of simulation without worrying about convergence

issues.

Fig. 4 (a) Basic π model for metal trace on lossy substrate; (b) The equivalent circuit model of vertical crosstalk between clock channel and CRWs.

Fig. 5. Comparison of simulated crosstalk between EM solution and equivalent model for (a) S_{11} , (b) S_{33} , (c) S_{31} , (d) S_{41} from 0.1 GHz to 40 GHz.

TABLE II
PARAMETERS IN EQUIVALENT CIRCUIT MODEL FROM 0.1GHZ TO 40GHZ.

$R_{s1}(\Omega)$	$R_{s2}(\Omega)$	$R_{c1}(\Omega)$	$R_{c2}(\Omega)$	L_S (pH)	L_C (pH)	L_{signal} (pH)	L_{clock} (pH)	C_{bond} (fF)
4.1	166.17	0.046	53.94	37.9	89.1	442	142.4	1.96
C_{IMD1} (fF)	C_{ss} (fF)	C_{signal} (fF)	C_{clock} (fF)	C_{sub} (fF)	R_{sub} (Ω)	M_{CSa} (pH)	M_{CSb} (pH)	M_{SS} (pH)
30.4	12.7	2.49	15.6	8.25	1690	140	140	406.64

3.Impacts of Design Variables on Vertical Noise Coupling

This section studies the impact of design variables on vertical noise coupling including geometry dimension, different stacking processes, and layout floorplan. Moreover, four methods which can reduce the crosstalk between clock channel and CRWs are discussed and compared.

3.1. Geometry Dimension

To investigate the influence of dimension on crosstalk, the length and width of metal traces (i.e. clock channel and CRWs) are varied separately. Fig. 6(a) plots the S_{31} from 0.1 GHz to 100 GHz for different metal length of 600 μ m and 1200 μ m, and Fig. 6(b)

depicts the crosstalk simulated at 3 GHz when the metal length varies from 500 μm to 1200 μm . As expected, the crosstalk increases by 5 dB for both NEXT and FEXT when metal length doubles since the mutual inductance is proportional to metal length. While the FEXT is 2 dB lower than the NEXT under the same length. Besides, the self-resonance frequency is also decreased for longer metal due to larger self-inductance. Fig. 7 plots the impact of metal width on crosstalk. Since the capacitive coupling is proportional to metal width, while the inductive coupling is a weak function of metal width, and the inductive coupling dominates below 15 GHz as described in section 2.3, the resulting crosstalk is almost unchanged for different metal widths below 15 GHz. However, the simulated crosstalk has an observable difference at frequency above 15 GHz because the capacitive coupling starts to manifest itself. As for the impact of TSV parameters, the radius of TSVs as well as the thickness of insulator have almost no impact on crosstalk. However, the depth (length) of TSV which is directly related to the thickness of top substrate does have an impact on vertical noise coupling or crosstalk, because both the inductive coupling and capacitive coupling become stronger as the TSV depth reduces.

Fig. 6 (a) Comparison of S_{31} with different clock channel and CRW length from 0.1 GHz to 100 GHz. (b) Impact of metal length on crosstalk at 3 GHz.

Fig. 7. (a) Comparison of S_{31} with different metal width from 0.1 GHz to 100 GHz. (b) Impact of width on crosstalk at 3 GHz.

3.2. 3D Stacking Process

The 3D stacking process will also have effect on the crosstalk, such as substrate thickness, face-to-face stacking or face-to-back stacking. Before stacking the analog chip on top of the digital chip, wafer thinning is required to reduce the thickness of top substrate and reveals the bottom of TSVs for micro-bump soldering. The impact of top substrate thickness on crosstalk is illustrated in Fig. 8, where a thinner substrate results in larger crosstalk since shorter coupling path leads to stronger coupling. As we know, the thickness of the substrate is related to TSV depth, which is restricted by the achievable aspect ratio (ratio of depth to diameter) from micro-fabrication process, so there is a trade-off between the crosstalk and substrate thickness or TSV depth.

Fig. 8 (a) Comparison of S_{31} with different thickness of top substrate from 0.1 GHz to 100 GHz. (b) Impact of top substrate thickness on crosstalk at 3 GHz.

Fig. 9 Comparison of S_{31} with different bonding process (face-to-face and face-to-back).

In Fig. 9, face-to-face (bottom tier face bonded to top tier face) and face-to-back (bottom tier face bonded to top tier back) bonding processes are compared for crosstalk performance. As expected, coupling noise in face-to-face bonding is worse than face-to-back bonding due to the short distance between clock channel and CRWs.

3.3. Layout Floorplan

The baseline model is constructed as the two CRWs located right above the clock channel in parallel with zero offset along x-direction. But in reality, the clock channel may not be in parallel with the CRWs, moreover, even they are in parallel, it is very likely that they offset from each other along the x-direction. Fig. 10 reports the crosstalk from 0.1 GHz to 100 GHz for different wire direction, i.e. clock channel is parallel or perpendicular to CRWs. The result clearly shows the crosstalk in perpendicular case is dramatically decreased compared to the parallel case, because both inductive and capacitive coupling are reduced. Specifically, the inductive coupling is reduced since the magnetic field excited by current flowing through clock channel and CRWs are within two orthogonal planes, and the capacitive coupling is also reduced due to smaller overlap area. Fig.11(a) compares the crosstalk when the horizontal offset (Δx) is 0, 40 μm and 80 μm . The result indicates that a larger offset leads to a smaller crosstalk, since the distance between clock channel and CRWs is increased. From Fig. 11(b), it is also shown that the offset of two CRWs induces difference between S_{31} and S_{51} , as the CRW 'a' and 'b' are no longer symmetrical with respect to clock channel.

Fig. 10 Comparison of noise transfer functions with different wire direction from 0.1 GHz to 100 GHz.

Fig. 11 (a) the comparison of S_{31} for different offset between clock channel and CRWs; (b) the crosstalk at the port of CRW 'a' and 'b' under 80μm offset.

3.4. Methods for Crosstalk Reduction

This section suggests four possible ways to reduce the vertical noise coupling between the analog and digital chips. The first method is to route the clock channel in perpendicular to the CRWs as discussed in section 3.3, which will greatly reduce the crosstalk. The second way is adding a metal ground plane at the bottom of top substrate to shield the electromagnetic field. As depicted in Fig. 12(a), the value of S_{31} with 1 μm thick copper ground plane are 18 dB lower than the baseline model without ground plane below self-resonant frequency. Another way to isolate the two tiers is to make the ground grids closer to the CRWs, which can reduce the mutual inductance and hence the inductive coupling between clock channel and CRWs. As a comparable study, the space between ground grids and CRWs is reduced from 190 μm (baseline model) to 30 μm and the simulated S_{31} is shown in Fig. 12(b) for both cases. The result tells that the crosstalk is much smaller for reduced distance between CRWs and ground grids, which also proves that the inductive coupling is the main coupling mechanism. The last suggested way to improve noise immunity is to make the two CRWs 'a' and 'b' differential so as to cancel the common-mode noise from the digital clock. Fig. 13 compares four different methods for crosstalk reduction, which demonstrates that using differential CRWs is the most effective approach for reducing crosstalk. However, it should be noted that the simulation is based on the ideal case with no horizontal offset between CRWs and clock channel. In other words, the CRW 'a' and 'b' are symmetrical with respect to the clock channel, therefore the noise coupling can be cancelled out when subtracting 'b' from 'a'. And the crosstalk cannot be completely cancelled out if a finite offset exists, as shown in Fig.11(b), which is the most likely scenario in real chip implementation. Besides, using differential CRWs in analog circuits requires more chip area and also consumes more power.

Fig. 12. (a) Comparison of S_{31} with and without ground plane. (b) Comparison of S_{31} for different space between CRWs and ground grids.

Fig. 13. Comparison of different methods for crosstalk reduction.

4. Time Domain Analysis

In this section, to analyze the transient noise coupling characteristics in time domain, the equivalent circuit model in Fig. 4(b) is imported into Advanced Design System (ADS). CMOS inverters are used as drivers and loads at each terminal of the clock channel and CRWs. To ensure the driving capability, the channel length and width of the CMOS inverter driving the clock channel at Port 1 is chosen as 100 nm and 240 μm respectively. While all the other MOSFETs' channel length and width are set to 100 nm and 5 μm. During simulation, the DC bias voltage on CRWs 'a' and 'b' is kept zero, and the input of inverter connected to clock channel is driven by a square wave with 1.2V amplitude, which is the supply voltage for 90 nm CMOS technology.

Fig. 14(a)-(f) illustrate the noise voltage measured at the near-end (Port 3) and far-end (Port 4) of CRW when a 1.2 V clock signal is applied to clock channel with different frequency and rising/falling time. Fig.14(a) plots the transient response to a 0.1 GHz clock signal with 1 ns rising/falling edge (10% of clock period), where the NEXT exhibits a 6.8 mV square-shape noise with sharp peaks (11 mV_{pp}) at its rising and falling edge. The square-shape noise is coupled from the 0.1GHz clock signal, and the peaks are generated by the sharp rising/falling edge of clock signal. Fig.14(b) plots the FEXT with the similar square-shape noise as NEXT but larger peaks (23 mV_{pp}) near the rising/falling edge. This is because the FEXT is transmitting in forward direction and grows larger with longer coupling length [22]. If the rising and falling time is reduced to 100 ps while keeping the same clock period (0.1 GHz), both NEXT and FEXT exhibit ringing due to faster clock switching, and the peak-to-peak noise voltage increases from 11 mV_{pp} to 52 mV_{pp} for NEXT and from 23 mV_{pp} to 173 mV_{pp} for FEXT, respectively, as shown in Fig. 14(c)-(d). Note that such large noise amplitude will bring nonnegligible influence on analog circuit. When clock frequency is increased to 1GHz while maintaining the same rising/falling edge (100 ps), the NEXT and FEXT are shown in Fig.14(e)-(f), which also have ringing effect, and the peak-to-peak value of the coupled noise remain the same, since they have same rising/falling time.

To further study the noise transfer characteristics, the Fast Fourier Transformation (FFT) is applied to coupled noise signals in Fig. 14(a)-(f). As depicted in Fig. 15(a)-(b), the noise spectrum clearly shows frequency components at 0.1GHz, 0.3GHz and 0.5GHz, which are the fundamental tone, 3rd and 5th harmonics of the square-shape noise waveform in Fig. 14(a)-(b). However,

when the rising/falling time reduces from 1 ns to 100 ps, the corresponding spectrum shown in Fig. 15(c)-(d) exhibits more tones besides the harmonics of the clock frequency (0.1 GHz), because the peak-to-peak noise voltage at the rising/falling edge increases dramatically in time domain. Finally, Fig 15(e)-(f) plots the FFT results of the coupled noise from 1 GHz clock with 100 ps rising/falling time, which have frequency components up to around 25 GHz. Finally, by comparing the noise spectrum of NEXT with that of FEXT from Fig.15(a)-(f), it can be concluded that the FEXT is more severe than the NEXT, which is consistent with the time domain results shown in Fig.14(a)-(f).

Fig. 14. Simulated NEXT (Port 3) and FEXT (Port 4) of the baseline model in time domain. (a) NEXT from a 0.1GHz square wave clock signal with 1ns rising/falling time, (b) FEXT from a 0.1GHz square wave clock signal with 1ns rising/falling time, (c) NEXT from a 0.1GHz square wave clock signal with 100ps rising/falling time, (d) FEXT from a 0.1GHz square wave clock signal with 100ps rising/falling time, (e) NEXT from a 1GHz square wave clock signal with 100ps rising/falling time, (f) FEXT from a 1GHz square wave clock signal with 100ps rising/falling time.

Fig. 15. Fast Fourier Transformation (FFT) of the noise waveform shown in Fig.14(a)-(f).

5. Conclusions

This paper analyzes the vertical noise coupling between clock channel and CRWs in TSV-based 3D mixed-signal integration. Full-wave EM model and broadband equivalent circuit model are built and the simulation results show a good agreement. Coupling mechanism is explored based on transfer impedance, which demonstrates a pure inductive characteristics from 0.1 GHz to 15 GHz, while the capacitive coupling manifests itself at frequency above 15 GHz. The effect of design variables on the crosstalk is also discussed, such as geometry dimension, different stacking processes, and layout floorplan. Four approaches for improving the noise immunity between clock channel and CRWs are suggested and compared, including keeping CRWs closer to ground grids, clock channel being perpendicular to CRWs, adding a ground plane, and applying differential CRWs. In the end, the noise coupling is simulated in time domain for different clock frequency and rising/falling time, which shows the peak-to-peak noise voltage increases with shorter rising/falling time and the far-end noise magnitude is much greater than the near-end.

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Graphical Abstract