Design of microstrip ultrafast pulses interconnect for THz-analog and digital applications

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Abstract

In this article, the propagation characteristic of transient pulses over the designed microstrip THz interconnects by various conducting materials is analyzed. The Fourier inverse transformation approach is used to visualize the effect of propagation of transient pulses over THz interconnects made by the various conducting materials which are widely used in Complementary Metal Oxide Semiconductor (CMOS) technology named Copper, Gold, Aluminum, Tungsten, Tungsten-Silicide (W-Si₂) and Poly-Si respectively. All parameters of transient pulse like rising time, fall time, propagation delay, settling time, overshoot, undershoot and ringing, etc. have calculated using the proposed circuit model and it has been verified with Finite Elements Method based EM-Simulation results. The distortion due to geometrical dispersion is significant; while a loss could be an important factor depending on the dielectric loss of the substrate. This information could be useful in designing both on board and on-chip interconnects. The proposed circuit model of microstrip THz interconnects and algorithmic frameworks proved the usefulness of rapid design, which is extremely challenging when using conventional methods. To the awareness, this is the attempt to efficient solving of this type of design problems; especially in the context of explicit THz interconnect design for CMOS technology.

Keywords THz interconnect · Dispersion · Distortion · Rise-time · Settling time · Ringing · Delay time · Gaussian · Modulated · Exponential signal · Circuit model · Integrated model

1 Introduction

The propagation of ultrafast pico-second pulse on the planar transmission line has been extensively studied. Ultra-fast pulses are useful to the modern fast computers, communication, radar, microwave and millimeter wave generation, study of material properties, time-domain network analyzer, Very Large Scale Integration (VLSI) interconnect technology [1–6] etc. Microstrip line is most popular planar line for both analog microwave applications and fast digital applications. The distortion in the electrical pulses traveling on the microstrip line is due to the geometrical dispersion, material dispersion, conductor loss, and dielectric loss. The distortion due to geometrical

dispersion is significant; while a loss could be an important factor depending on the dielectric loss of the substrate. The conductor loss role is usually of secondary nature and for good conductor line, its role could not be ignored. The distortion in pulses on the microstrip line due to the geometrical dispersion has been analyzed by using the Finite Element Method (FEM) based HFSS (High-Frequency Structure Simulator) EM-simulator, spectral domain analysis and closed-form model [7].

The effect of conductor and dielectric loss through closed-form models has been included in amplitude factor and phase-factor. Apart from distortion in the digital pulses, propagating on the microstrip line useful for the Microwave Integrated Circuits/Monolithic Microwave

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Integrated Circuits (MIC/MMIC), distortion in Gaussian pulses has also been investigated. It is also useful to estimate the bandwidth of the microstrip line under various conditions [8–12]. The pulse shaping of the exponential pulse has also been investigated [13].

Pulse distortion on microstrip line Several parameters are used to quantify distortion in a rectangular pulse. Distortion in the rectangular pulse is very important in this study. Therefore, these parameters are given below in order to avoid ambiguity in our time-domain discussion and also describe reasons for the distortion in a rectangular pulse [14–23].

Quantification of distortion in pulse Fig. 1 shows the figures of merit with respect to step response of any linear network. The figures of merits are the parameters used to quantify distortion in a rectangular pulse as follows [24].

Rise-time is related to the 3 dB frequency f_{3dB} and knee frequency f_{knee} as $0.35/T_r$ and $0.5/T_r$ respectively. We note that $f_{knee} > f_{3dB}$. The f_{3dB} bandwidth is defined as the passage of only 50% of the input power to the output. It is not a useful description of the bandwidth of interconnect. The knee frequency f_{knee} bandwidth permits passage of most energy in digital pulse below f_{knee} . Thus 0.2 ps rise time requires 250 GHz bandwidth microstrip line in order to avoid distortion due to bandwidth limitation [25–27].

A very short interconnect described by the RC-circuit, there is no ringing and overshoot. The amplitude of the pulse response goes negative and it requires settling time. It is expressed as a percentage of the final value [28].

A distortion-less line provides only the time shifting of the input pulse. A pulse transmitting line requires the output sequence of pulse have almost the same distinct pulse shapes without any so-called inter-symbol interference. It is achieved a short rise-time and short settling-time of the line and the long rise-time and long settling-time results in the smearing of pulses shown in Fig. 2 respectively [29].

Fig. 1 Figures of merits to quantify distortion in a rectangular pulse

Fig. 2 Non-smeared and smeared pulses due to the rise-time and settling-time

Factors causing distortion in pulse propagation The distortion in pulse propagation on the microstrip line is due to the following factors:

Losses Conductor loss, dielectric loss, and radiation loss, etc. **Mode coupling:** coupling of higher order modes to the fundamental mode. **Dispersion:** geometrical dispersion and material dispersion. This article concentrates on pulse distortion due to losses and dispersion in the microstrip line [30–36]. The dispersion in line gives frequency dependent phase velocity of the spectral components of pulse on the line causing distortion in signals. Likewise, the frequency dependent losses also cause distortion in pulses [37]. The dispersion itself causes distortion by four mechanisms: (1) Distortion due to the

Fig. 3 Nature of distortion in digital pulse

Table 1 Input signals and their Fourier transforms, used for analysis

Signal's type	Time domain	Fourier transforms	Shape
Ideal square pulse	$V(t, z=0) = \begin{cases} A & -\frac{\tau}{2} \leq t \leq \frac{\tau}{2} \\ 0 & \text{otherwise} \end{cases}$	$V(\omega, z=0) = A\tau \frac{\sin\left(\frac{\tau}{2}\omega\right)}{\omega \frac{\tau}{2}}$
Equal rise/fall pulse	$V(t, z=0) = \begin{cases} 1 & 0 < t < T_1 \\ 1 - a(t - T_1)^2 & T_1 < t < T_1 + q\tau_1 \\ bt + e & T_1 + q\tau_1 < t < \tau - q\tau_1 \\ a(t - \tau)^2 & \tau - q\tau_1 < t < \tau \\ 0 & t > \tau \end{cases}$	$V(\omega, z=0) = \frac{8}{q(1-q)\tau_1^2} \times \frac{\sin\left[\left(\tau - \frac{\tau_1}{2}\right)\omega\right] \sin\left[\frac{q\tau_1\omega}{2}\right] \sin\left[\frac{(1-q)\tau_1\omega}{2}\right]}{\omega^3}$
Unequal rise/fall pulse	$V^*(t, z=0) = \begin{cases} 1 & 0 \leq t \leq \tau_1 \\ 0 & \text{otherwise} \end{cases}$	$V(\omega, z=0) = \left[\frac{1 - \exp(-j\omega\tau_1)}{(j\omega)} \right] \times \left[\frac{1}{(j\omega\tau_2 - \omega^2/\omega_1^2 + 1)} \right]$ where, $\omega_1^2 = \frac{1}{LC}$, τ_1 = Input pulse width $\tau_2 = \frac{L}{Z_0}$
Modulated square pulse	$V(t, z=0) = \begin{cases} A \cos\omega_0 t & -L \leq t \leq L \\ 0 & \text{otherwise} \end{cases}$	$V(\omega, z=0) = A \left[\frac{\sin\left[\frac{\tau}{2}(\omega - \omega_0)\right]}{(\omega - \omega_0)} + \frac{\sin\left[\frac{\tau}{2}(\omega + \omega_0)\right]}{(\omega + \omega_0)} \right]$
Ideal Gaussian pulse	$V(t, z=0) = A e^{-\frac{4t^2}{\tau^2}}$	$V(\omega, z=0) = \frac{A\sqrt{\pi}}{a} \exp\left(-\frac{\omega^2}{4a^2}\right) \text{ where } a = \frac{2}{\tau}$
Exponential pulse	$V^*(t, z=0) = e^{-\frac{t}{\tau_1}}$	$V(\omega, z=0) = \left[\frac{1}{(j\omega + 1/\tau_1)} \right] \times \left[\frac{1}{(j\omega\tau_2 + \tau_2/\tau_1 + 1)} \right]$ where, $\tau_1 = Z_0 C$, $\tau_2 = RC$

Fig. 4 Circuit diagrams generating unequal **a** rise-time and fall-time signal rectangular pulse and **b** exponential pulse

bandwidth limitation of microstrip transmission line (2) Distortion due to the pulse catching up phenomenon within the pulse-width (3) Distortion due to separation of the low-frequency component of the pulse (4) Distortion due to the separation of the high-frequency pulse.

If a small portion of high-frequency components is separated from the main pulse response of line will look

like the response of the low-pass filter. Whereas, if a small portion of low-frequency components is separated from the main pulse, the response of the line will look like a response from high-pass filter [38]. Such distortions in the rectangular pulse are shown in Fig. 3a. Figure 3b shows an increase in rise-time i.e. longer rise-time due to the absence of high-frequency spectral components. It is due to the limited bandwidth of a line. Figure 3c shows tilt in the pulse due to the separation of low-frequency component from the main pulse. Figure 3d shows distortion in the pulse due to the separation of low-frequency components combined with separation of high-frequency spectral components. Figure 3e shows adding up of high-frequency components during the pulse width time. It is due to delay in the propagation of high-frequency components of the pulse. It results in ringing at the leading and trailing edges of the pulse due to the overemphasis of high-frequency components [39–42].

The dispersion causes broadening of Gaussian pulse i.e. increase in Full width at half maximum (FWHM) at the output of line and spreading of ringing in the time-domain. This is due to the redistribution of spectral energy [43–46]. Such temporal spreading of the pulse increases the separation of the next pulse, thereby decreasing the bit rate of the line. So for the high-speed pulse propagation line

should be less dispersive. If we concentrate on the main portion of the Gaussian pulse the increase in its pulse width i.e. FWHM is an indication of separation of high-frequency components of the original pulse. The increase in FWHM is a measure of dispersive nature of the line. In the filter theory, such absence of high-frequency components is an indication of band limitation of the filter. Thus the Gaussian pulse-width i.e. FWHM is related to the bandwidth of line [14, 47–52].

In addition, an aerosol-jet printing technique is realized an optical transmitter assembly of a microring modulator and a high-speed CMOS driver in which the bonding wires are replaced by aerosol-jet printed silver interconnects [53]. Furthermore, conventional complementary metal oxide semiconductor (CMOS)-based electronic interconnects are not expected to keep up with the envisioned future board-to-board and chip-to-chip (within multi-chip-modules) interconnect requirements because of bandwidth-density and power-consumption limitations [54]. However, with the CMOS technology shrinking, it becomes critical to delivering high-quality global clock signal with low propagation delay and hence conventional metallic interconnect seems to meet its bottleneck, as a clock distribution network (CDN) might consume up to 50% of the overall power [55]. So, the Propose formation of ultra-low-cost visible-light integrated optical links by interfacing dense micro-pixelated LED arrays with matching multi-layered multimode polymer waveguide arrays [56].

This paper presents, a detailed study on the effect of various physical parameters of the microstrip transmission line on digital pulses using the circuit model. The accuracy of the circuit model to analyze the pulse shape has been tested against the full-wave analysis & provides almost identical results in time-domain for various parameters of a lossy microstrip line. We also compare the results of the integrated model, that accounts for the effect of the loss factor only on the amplitude terms of the spectral components and dispersion are accounted for in the phase-factor.

In the case of the low-loss microstrip, both the circuit and integrated models provide almost identical results. However, for highly lossy microstrip their results differ significantly. Thus under such circumstance, the direct model is not adequate and we should use the circuit model for more realistic analysis [7]. Finally, we have investigated the effect of various physical parameters of a microstrip line on the pulse defining parameters- rise time, fall-time, propagation delay, settling time, overshoot and undershoot, etc. This information could be useful for designing both on board and on-chip interconnects. MATLAB programming and Ansoft HFSS simulation tools are used for the entire analysis.

2 Analysis of signals

A dispersive and lossy transmission line gives frequency dependent losses and frequency dependent phase velocity to each spectral components of the input time-domain signal. Thus with the help of Fourier transform input time domain signal is decomposed into its spectral components in the frequency domain.

Numerical method to compute response Various voltage $v(t, z = 0)$ signals in time-domain used as the input signals to the microstrip line, along with their Fourier transforms are shown in Table 1.

The Fourier transform of the time-domain signal $v(t)$ is obtained from the following expression

$$V(\omega, z = 0) = \int_{-\infty}^{\infty} v(t, z = 0) e^{-j\omega t} dt \quad (1)$$

The unequal rise-time and fall-time signal and the exponential signal marked with (*) in Table 1 are generated through LC and RC circuit shown below in Fig. 4. The Fourier transforms shown in Table 1 for these pulses are for the time-domain signals at the out of the circuits [46].

Fig. 5 **a** Effect of ξ (KI) on overshoot of 50 ps digital pulse. Complete input pulse with overshoot and undershoot **b** effect of ξ (KI) on overshoot of 50 ps digital pulse enlarged the view of overshoot

Table 2 Effect of ξ (KI) on Overshoot

ξ (KI)	100	200	300	400	500	600	700	800	900	1000
Rise time (ps)	1.3	1.3	1.3	1.3	1.3	1.3	1.3	1.3	1.3	1.3
Overshoot (%)	6.0	6.2	-2.0	2.0	3.0	-0.5	-0.5	3.0	0.5	1.3

Fig. 6 Comparisons of distortion in output pulses as computed by HFSS curve-fitted model and circuit model. For square pulse $\tau = 50$ ps and for Gaussian pulse FWHM = 10 ps, $w/h = 0.1$

These output signals are fed to the loss microstrip line. The time-constants controlling shape of the pulses is defined in Table 1. The Fourier transform of the pulse provides spectral components. Thus at the end of line length without any reflection the frequency domain response spectral components are given by

$$V(\omega, z = L) = H(\omega) \times V(\omega, z = 0) \quad (2)$$

where $H(\omega)$ is the transfer function of the matched line. For the lossy transmission line, it is given by

$$H(\omega) = e^{-\gamma(\omega)L} \quad (3)$$

where frequency dependent complex propagation constant $\gamma(\omega)$ is given by

$$\gamma(\omega) = \alpha_T(\omega) + j\beta(\omega) \quad (4)$$

$\beta(\omega)$ is the frequency dependent phase constant of the lossy microstrip line and $\alpha(\omega)$ is the total attenuation

factor of the microstrip line due to dielectric and conductor losses. The frequency dependent propagation is obtained from the integrated model using a combination of the MKJ-dispersion model and LDM [7]. It is also determined by the circuit model that accounts for the effect of losses on the propagation constant. Furthermore, it is also determined by the curve-fitted model based on the data obtained from HFSS. This model also accounts for the influence of losses on the propagation constant. The total attenuation factor is

$$\alpha_T(\omega) = \alpha_d(\omega) + \alpha_c(\omega) \quad (5)$$

where $\alpha_d(\omega)$ and $\alpha_c(\omega)$ are attenuation factor due to dielectric and conductor losses respectively. The $\alpha_d(\omega)$ and $\alpha_c(\omega)$ are also obtained from independent closed-form models forming parts of the integrated model. However, the losses are to some extent also influenced by the propagation constants. This is again accounted for by the circuit

Fig. 7 Comparisons of distortion in output pulses as computed by HFSS curve-fitted model and circuit model. For square pulse $\tau = 50$ ps and for Gaussian pulse FWHM = 50 ps, $w/h = 1.0$

model and the curve-fitted model total attenuation factor $\alpha_T(\omega)$ using HFSS data. All these models are discussed in [7]. Finally, the pulse in the time domain at end of the line length $z = L$ is obtained from the inverse Fourier transform

$$v(t, z = L) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H(\omega) \times v(\omega, z = 0) e^{j\omega t} dt \quad (6)$$

The complexity of the frequency-dependent phase constant $\beta(\omega)$ does not permit solving the Eq. (6) in the closed-form. Numerical integration is the most straight forward technique for evaluating Eq. (6). However, its accuracy depends on the amount of computer time and storage space available. Taylor series expansion method could be used for modulated RF pulses. Taylor series expansion method [29] is an approximation to full integration of the Eq. (6). It is slightly less accurate than numerical integration, but it requires much less computer time to evaluate the integral, especially for RF pulses. The FFT process has also been used. In Eq. (6), the limits of integration are $-\infty < \omega < \infty$. However, beyond a certain radian frequency ω_L the contribution of frequency component to the integral

are negligible. Narrower pulses have higher frequency content. This leads to a higher value of ω_L . If τ is the width of the pulse, then $\omega_L = \xi/\tau$, Where, ξ is a constant which depends on the wave-shape.

Its value is determined through numerical experimentation. Thus limits of Eq. (7) are restricted to $\pm \omega_L$

$$v(t, z = L) = \frac{1}{2\pi} \int_{-\omega_L}^{\omega_L} v(\omega, z = 0) e^{-\alpha(\omega)L} \cdot e^{j[\omega t - \beta(\omega)L]} d\omega \quad (7)$$

The voltage at the output of line is approximated by the trapezoidal rule [16]. For the lossless line, it is given by

$$v(t, z = L) = \frac{1}{2\pi} \sum_{i=1}^N v(\omega_i, z = 0) \cdot e^{j[\omega_i t - \beta(\omega_i)L]} \Delta\omega_i \quad (8)$$

For the lossy line voltage at the output is

$$v(t, z = L) = \frac{1}{2\pi} \sum_{i=1}^N v(\omega_i, z = 0) e^{-\alpha(\omega_i)L} \cdot e^{j[\omega_i t - \beta(\omega_i)L]} \Delta\omega_i \quad (9)$$

Fig. 8 Comparisons of distortion in output pulses as computed by HFSS curve-fitted model and circuit model. For square pulse $\tau = 50$ ps and for Gaussian pulse FWHM = 50p, $w/h = 10$

where N is the number of division in the frequency spectrum and $\Delta\omega = 2\xi/\tau N$ is the width of each uniform segment. Since we are concerned only with the real part of the pulse, Eq. (9) becomes

$$v(t, z = L) = \frac{1}{2\pi} \sum_{i=1}^N v(\omega_i, z = 0) e^{-\alpha(\omega_i)L} \cdot \cos(\omega_i t - \beta(\omega_i)L) \Delta\omega_i \quad (10)$$

For the lossless line, we have $\alpha(\omega) = 0$.

Numerical Experimentation to Determine ξ (KI) for Digital Pulse In above equation $V(\omega_i, z = 0)$ is the Fourier transform of the digital pulse given in Table 1. In time-domain it is a perfect rectangular pulse. However using Eq. (10), we can regenerate the 50 ps pulse at input, $z = 0$. Figure 5 shows such regenerated pulse by changing the value of ξ from 100 to 1000.

Figure 5a that even the input pulse is no longer a perfect pulse with zero rise-time and with no overshoot and no undershoot. The input 50 ps pulse has rise-time 1.3 ps. It requires 385 GHz bandwidth lossless microstrip line to support its undistorted propagation. Figure 5b shows enlarged overshoots that are controlled by the value of ξ (KI). The results of experimentation are summarized in Table 2.

The rise-time is not affected by the value of ξ . However, % overshoot is controlled by ξ and its suitable value is 600–700. For different pulse-width, its value could change. Balanis et al. [30] have suggested its value 500 for the digital pulse and 20 for the Gaussian pulse, as the rise time of a Gaussian pulse is slower.

Separation of frequency components We have mentioned above that the tilt in the response, shown in Fig. 3c, is due to the poor low-frequency response of the line. While pulse propagates on the line, its low-frequency spectral component gets separated from the main pulse resulting into the tilt. The separation frequency f_L can be estimated as follows:

The phase velocity of pulse at separation frequency f_L is related to length L and propagation delay time (t_d) as $V_p(f_L) = (L/t_d)$. The delay time t_d is the time of arrival of the digital pulse which is measured at 50% of the amplitude of the pulse. In the above equation, L is known and t_d is obtained from the response computed from the above discussed numerical method. Thus, the effective permittivity of microstrip line K at frequency f_L is computed from the expression $K = \{V_0/(V_p \times f_L)\}^2$. Where V_0 is the velocity of EM-wave in free-space. As K is now known, the separation frequency f_L can be obtained by solving the

Table 3 Comparison static, circuit and curve-fitted HFSS models for time-domain parameters for microstrip line $w/h = 0.1$, substrate with $\epsilon_r = 11.9$, $h = 0.45$ mm, $t/h = 0.05$, $\tan\delta = 0.001$, strip conductivity $\sigma_0 = 3.8 \times 10^7$ S/m, $L = 3.0$ mm

Models response	Pulse-width (ps) 50%	Rise-time (ps) 10–90%	Delay-time (ps) 0–50%	Settling-time (ps) $\pm 2\%$ of flat amplitude	Over-shoot % of flat amplitude	Under-shoot % of flat amplitude	Fall-time (ps) 90–10%	Amplitude (V)	Pulse extension out of static response		
									Before (ps)	After (ps)	
Input pulse: square with negligible rise-time, pulse width $\tau = 50$ ps [30–34]											
Static	50	0.8	0.5	3.0	8.0	8.0	0.8	1	0	0	
Circuit model	50	10.0	14.5	27.0	35.5	50.0	10.5	0.974	13	7	
Curve-fitted HFSS	50	10.0	14.5	27.0	35.5	50.0	10.5	0.974	13	7	
Input pulse: square with finite equal rise-time and fall-time ($t_r = 20$ ps), pulse-width $\tau = 50$ ps, $q = 0.12$ [36]											
Static	50	14.0	10.0	20.0	0.0	0.0	14.0	1	0	0	
Circuit model	50	15.5	15.5	33.0	6.5	6.5	16.0	0.94	7	6	
Curve-fitted HFSS	50	15.5	15.5	33.0	6.5	6.5	16.0	0.94	7	6	
Input pulse: square with unequal Rise time and fall time $\tau_1 = 50$ ps, $\tau_2 = 5$ ps, $\omega_1 = 10^{12}$ Hz [47]											
Static	50	11.5	8.0	30.0	0.0	0.0	10.5	1	0	0	
Circuit model	50	10.0	11.5	20.0	1.5	0.5	9.5	0.93	0	0	
Curve-fitted HFSS	50	10.0	11.5	20.0	1.5	0.5	9.5	0.93	0	0	
Input pulse: FWHM = 10 ps [30–34]											
Static	10	8.0	8.0	–	0.0	0.0	8.0	1	0	0	
Circuit model	10	10.0	14.5	–	0.0	16.0	9.5	0.89	10	0	
Curve-fitted HFSS	10	10.0	14.5	–	0.0	16.0	9.5	0.89	10	0	

Table 4 Comparison static, circuit and curve-fitted HFSS models for time-domain parameters for microstrip line $w/h = 1.0$, substrate with $\epsilon_r = 11.9$, $h = 0.45$ mm, $t/h = 0.05$, $\tan\delta = 0.001$, strip Conductivity $\sigma_0 = 3.8 \times 10^7$ S/m, $L = 3.0$ mm

Models response	Pulse-width (ps) 50%	Rise-time (ps) 10–90%	Delay-time (ps) 0–50%	Settling-time (ps) ±2% of flat amplitude	Over-shoot % of flat amplitude	Under-shoot % of flat amplitude	Fall-time (ps) 90–10%	Amplitude (V)	Pulse extension out of static response	
									Before (ps)	After (ps)
Input pulse: square with negligible rise-time, pulse-width $\tau = 50$ ps										
Static	50	0.8	4.5	8.0	1.3	1.4	0.8	1	0	0
Circuit model	50	10.8	17.5	32.0	40.5	64.0	10.5	0.96	13	7
Curve-fitted HFSS	50	10.8	17.5	32.0	40.5	64.0	10.5	0.96	13	7
Input pulse: square with finite equal rise-time and fall-time ($t_r = 20$ ps), pulse-width $\tau = 50$, $q = 0.12$										
Static	50	14.0	10.0	19.0	0.0	0.0	14.0	1	0	0
Circuit model	50	14.5	17.5	39.0	6.7	8.3	16.0	0.97	7	6
Curve-fitted HFSS	50	14.5	17.5	39.0	6.7	8.3	16.0	0.97	7	6
Input pulse: square with unequal rise time and fall time $\tau_1 = 50$ ps, $\tau_2 = 5$ ps, $\omega_1 = 10^{12}$ Hz										
Static	50	12.0	6.5	25.0	0.0	0.0	10.0	1	0	0
Circuit model	50	8.5	4.5	14.0	0.2	0.0	9.5	0.97	0	0
Curve-fitted HFSS	50	8.5	4.5	14.0	0.2	0.0	9.5	0.97	0	0
Input pulse: FWHM = 10 ps										
Static	50	30.0	37.0	–	0.0	0.0	30.0	1	0	0
Circuit model	50	31.5	38.0	–	0.0	0.0	29.5	0.99	0	0
Curve-fitted HFSS	50	31.5	38.0	–	0.0	0.0	29.5	0.99	0	0

Table 5 Comparison static, circuit and curve-fitted HFSS models for time-domain parameters for microstrip line $w/h = 10.0$, substrate with $\epsilon_r = 11.9$, $h = 0.45$ mm, $t/h = 0.05$, $\tan\delta = 0.001$, strip Conductivity $\sigma_0 = 3.8 \times 10^7$ S/m, $L = 3$ mm

Models response	Pulse-width (ps) 50%	Rise-time (ps) 10–90%	Delay-time (ps) 0–50%	Settling-time (ps) ± 2% of flat amplitude	Over-shoot % of flat amplitude	Under-shoot % of flat amplitude	Fall-time (ps) 90–10%	Amplitude (volt)	Pulse extension out of static response	
									Before (ps)	After (ps)
Input pulse: square with negligible rise-time, pulse width $\tau = 50$ ps										
Static	50	0.8	5.5	10.0	3.1	3.2	0.8	1	0	0
Circuit model	50	2.5	30.5	50.0	11.5	11.6	4.5	0.97	26	4
Curve-fitted HFSS	50	2.5	30.5	50.0	11.5	11.6	4.5	0.97	26	4
Input pulse: square with finite equal rise-time and fall-time ($t_r = 20$ ps), pulse–width $\tau = 50$, $q = 0.12$										
Static	50	14.0	10.0	19.0	0.0	0.0	14.0	1	0	0
Circuit model	50	13.5	33.5	65.5	0.5	4.0	14.5	0.97	24	6
Curve-fitted HFSS	50	13.5	33.5	65.5	0.5	4.0	14.5	0.97	24	6
Input pulse: square with unequal rise time and fall time $\tau_1 = 50$ ps, $\tau_2 = 5$ ps, $\omega_1 = 10^{12}$ Hz										
Static	50	10.5	5.5	25.0	0.0	0.0	10.0	1	0	0
Circuit model	50	10.0	4.5	27.0	0.0	0.0	9.5	0.97	0	0
Curve-fitted HFSS	50	10.0	4.5	27.0	0.0	0.0	9.5	0.97	0	0
Input pulse: FWHM = 10 ps										
Static	50	30.0	37.0	–	0.0	0.0	30.0	1	0	0
Circuit model	50	31.5	38.0	–	0.0	0.0	29.5	0.99	0	0
Curve-fitted HFSS	50	31.5	38.0	–	0.0	0.0	29.5	0.99	0	0

Fig. 9 Distortion in 20 ps Gaussian pulse due to various factor on microstrip line **a** 5.0 inch **b** 20.0 inch

Fig. 10 50 ps square-wave on microstrip length 3.0 mm. Strip conductor Tungsten (1.82×10^7 S/m), resistivity of Si-substrate $\rho_s = \infty$, 100, $10 \Omega \text{ cm}$, $w = 10 \mu\text{m}$

Fig. 11 50 ps square-wave on microstrip line length 3.0 mm. Strip conductor tungsten (1.82×10^7 S/m), resistivity of Si-substrate $\rho_s = \infty$, 100, $10 \Omega \text{ cm}$, $w = 300 \mu\text{m}$

Fig. 12 50 ps square-wave on microstrip length 3.0 mm. Strip conductor Tungsten-Silicide (0.33×10^7 S/m) resistivity of Si-substrate $\rho_s = \infty$, 100, 10 Ω cm, $w = 10$ μ m

Fig. 13 50 ps square-wave on microstrip length 3 mm. Strip conductor tungsten-silicide (0.33×10^7 S/m) resistivity of Si-substrate $\rho_s = \infty$, 100, 10 Ω cm, $w = 300$ μ m

non-linear equation with the help of the MKJ-dispersion model, $\epsilon_{\text{reff}}(w, h, t, \epsilon_r, f) - K = 0$ [7]. Finally, the distortion in digital pulse due to the separation of low-frequency components is characterized by the tilt in square digital pulse as defined in Fig. 3c.

$$\text{Tilt} = \{(E_1 - E_2)/E\} \times 100\% \quad \text{where, } E = (E_1 + E_2)/2 \quad (11)$$

The separated high-frequency components are delayed and they combine with low-frequency components of the pulse during pulse width time, thus giving ringing at the leading edge due to the enhancement of high-frequency components. The ringing created due to the numerical process discussed above is known as Gibb's phenomenon.

3 Numerical results on pulse distortion

Integrated and circuit models used to compute the attenuation constants and the propagation constant of a lossy microstrip line [7]. The integrated model, also called the direct model as it's every section computes directly from individual expression dispersion, dielectric loss and conductor loss. The circuit model is a more accurate model for high loss microstrip line, as it considers the effect of losses on the propagation constant. We have also reported [7] curve-fitted HFSS models that are based on the data obtained from HFSS. Such curve-fitted models have to be obtained for each individual microstrip line. In this section first, we compare performances of both the curve-fitted

Fig. 14 50 ps square-wave on microstrip length 3 mm. Four types of strip conductor Resistivity of Si-substrate $\rho_s = \infty$, $w = 300 \mu\text{m}$, $t = 0.5 \mu\text{m}$

HFSS models and the circuit model in time-domain to obtain the time-domain response of the microstrip line. Next, we compare the performances of the circuit model and the integrated model (direct model) in time-domain. We also present design information for the effect of line width, length, strip conductor conductivity and substrate resistivity on the time-domain direct model parameters, such as rise-time, delay-time, overshoot, etc. of a lossy microstrip.

Performances of Models against HFSS The curve-fitted HFSS models are reported [7], for microstrip line with for various w/h as 0.1, 1.0, 10. The microstrip is considered on substrate with $\epsilon_r = 11.9$, $h = 0.45 \text{ mm}$, $\tan\delta = 0.001$ and strip thickness $t/h = 0.05$ are aluminum strip conductivity is $\sigma_0 = 3.8 \times 10^7 \text{ S/m}$. The curve-fitted HFSS models have a maximum deviation of 1.5% and an average deviation of 0.8% for the frequency dependent effective relative permittivity. It has an average deviation of 5.0% for the total loss computation.

These deviations are against the HFSS data over frequency range 0.01 GHz–1.0 THz. The circuit model has a max deviation of 4.0% and an average deviation of 2.5% for dispersion and the average deviation of 8.0% for total loss against the HFSS data over frequency range 0.01 GHz–1.0 THz. We may state that the circuit model has 1.7% deviation in the computation of dispersion with respect to the curve-fitted HFSS models and 3.0% deviation in the computation of total loss respect to the curve-fitted HFSS models [7]. We desire to see the effect of these deviations in the circuit models on performances of the pulse distortion with respect to the pulse distortion using curve-fitted HFSS models.

We have considered distortion in four types of input pulses—square pulse with negligible rise-time, finite equal rise-time and fall-time (t_r) square pulse, unequal rise-time and fall-time square pulse and Gaussian pulse. We have considered pulse-width $\tau = 50 \text{ ps}$ for square pulses and FWHM = 10 ps for Gaussian pulse. Figures 6, 7

Fig. 15 50 ps square-wave on microstrip length 3 mm. Four types of strip conductor Resistivity of Si-substrate $\rho_s = 100 \Omega \text{ cm}$, $w = 300 \mu\text{m}$

and 8 compares responses of these four input signals on $L = 3 \text{ mm}$ long microstrip of widths $w/h = 0.1, 1.0$ and 10 respectively. Microstrip line are low-loss lines as strip conductor is aluminum ($\sigma_0 = 3.8 \times 10^7 \text{ S/m}$) and loss-tangent is only 0.001 . Summary of investigation is shown in Tables 3, 4 and 5 for $w/h = 0.1, 1.0$ and 10.0 respectively. We note from responses and tables that both the integrated model and circuit modes give identical responses. Thus variations in a deviation between two models (1.7% in propagation constant and 3% in total loss) do not have an effect on the time domain response of rectangular and Gaussian pulses of 50 ps and 10 ps pulse width.

The extension of response after the end of the response of the static case shows that the pulse response spreads due to dispersion. It limits the speed of data transmission.

If we align the start of the dispersive pulse with the static response to get the physical response, the extension of the pulse due to dispersion will further increase, giving a more realistic picture on inter-symbol interference limiting the bit transmission rate. Our circuit model

accuracy with respect to time-domain response and their parameters have been confirmed. Now, we can check the accuracy of the integrated model (direct model) against the circuit model [7].

Comparison of direct model and circuit model In this section, we compare both the direct model i.e. the integrated model and the circuit model with respect to their pulse responses in time-domain. The direct model with changes in dispersion model and some variations in dielectric loss and conductor loss models has been used Balanis et al. [34]. They have not accounted for the effect of the strip conductor thickness in their dispersion models. As mentioned, earlier the direct model does not account for the effect of losses on the propagation constant. It is important for the high loss microstrip. We will examine both models for the lossless, low-loss and high-loss microstrip line.

In the case of the lossless microstrip, both models are identical and only dispersion effect is present. We consider the microstrip line taken by Balanis et al. [32]. This

Fig. 16 50 ps square-wave on microstrip length 3 mm. Four types of strip conductor resistivity of Si-substrate $\rho_s = 10 \Omega \text{ cm}$, $w = 300 \mu\text{m}$

microstrip has substrate with $\epsilon_r = 10.2$, $\tan\delta = 0.0015$, $h = 0.02 \text{ inch}$, $w = 0.001 \text{ inch}$, $t = 0.00067 \text{ inch}$, length $L = 20 \text{ inch}$, strip is conductor and ground is aluminum.

Another microstrip has length 5.0 inch. The Gaussian pulse has FWHM = 20 ps. The response of Gaussian pulse for both lines are shown in Fig. 9a, b for line length 5.0 inch and 20.0 inch respectively for various conditions- static lossless parameters, dispersion only with no loss, static case with conductor loss, static case with dielectric loss, dispersion and total loss i.e. the direct model and the circuit model.

We note from Fig. 9a that the lossless static case only shifts the pulse in time scale. For the short length line, dielectric loss is less as compared to the conductor loss. It is due to the low-loss substrate. The dispersion is the main reason for spreading of pulse, not losses. Both the direct model and circuit model provide an almost identical response. The models are non-causal and spreading

of the pulse is due to dispersion, the losses do not contribute towards it. They only reduce the amplitudes of the main pulse and riggings. This discussion is valid for longer line length shown in Fig. 9b. However, distortion is more. Both models provide identical response only for the low-loss line. It is not true for a high-loss microstrip. It is discussed below.

Figures 10 and 11 show the distortion in $\tau = 50 \text{ ps}$ square-wave pulses after 3.0 mm of propagation on Si substrate with $\epsilon_r = 11.9$, $h = 450 \mu\text{m}$. The strip conductor is made of tungsten ($1.82 \times 10^7 \text{ S/m}$) and its thickness is $t = 0.5 \mu\text{m}$. The resistivity of the substrate is considered- $\rho_s = \infty, 100, 10 \Omega \text{ cm}$. In Fig. 10 strip width w is $10 \mu\text{m}$, whereas for Fig. 11 it is $w = 300 \mu\text{m}$. Likewise, Figs. 12 and 13 show the distortion in $\tau = 50 \text{ ps}$ square-wave pulses after 3.0 mm propagation on Si-substrate for strip conductor made of tungsten-silicide ($0.33 \times 10^7 \text{ S/m}$). Other parameters are the same.

Fig. 17 50 ps square-wave on microstrip length 6.0 mm. Strip conductor W-Si (0.33×10^7 S/m) Resistivity of Si substrate $\rho_s = \infty, 100, 10 \Omega \text{ cm}$, $w = 10 \mu\text{m}$

We note that for resistivity of substrate $\rho_s \leq 100 \Omega \text{ cm}$, both models provide an identical response. However, for resistivity of substrate $\rho_s = 10 \Omega \text{ cm}$, responses of both models are different. Thus we have to use circuit-model for high-loss microstrip line. The effect of various parameters on distortion is discussed in the next section.

We have considered six materials as strip and ground conductors—Copper (5.8×10^7 S/m), Gold (4.1×10^7 S/m), Aluminum (3.0×10^7 S/m), Tungsten (1.82×10^7 S/m), Tungsten-Silicide (W-Si₂) (0.33×10^7 S/m), Poly-Si (0.02×10^7 S/m). We have examined several strip width $w = 1, 10, 100, 300, 450, 4500 \mu\text{m}$ and several lengths of microstrip $L = 0, 3, 6, 9, 12 \text{ mm}$.

Figures 14, 15 and 16 show distortion in $\tau = 50$ ps. Square-wave pulses after 3.0 mm of propagation on microstrip with conductors- Aluminum, Tungsten, Tungsten-Silicide (W-Si₂), Poly-Si. The pulse width 10 ps and 50 ps have been considered. We have studied distortion in pulses under these parameters. Some of the results are presented in this section. Three figures are for three Si substrates ($\epsilon_r = 11.9$, $h = 450 \mu\text{m}$) with resistivity $\rho_s = \infty, 100, 10 \Omega \text{ cm}$. Microstrip line has $w = 300 \mu\text{m}$, $t = 0.5 \mu\text{m}$. We note a gradual increase in attenuation and distortion in pulses with higher dielectric and conductor losses.

Figure 17 further presents distortion and attenuation in 50 ps square-wave on microstrip length having 6.0 mm.

The $10 \mu\text{m}$ wide microstrip is made of conductor W-Si (0.33×10^7 S/m) on three resistivities of Si-substrate $\rho_s = \infty, 100, 10 \Omega \text{ cm}$. First three results of Fig. 17 shows gradually increase in attenuation and distortion with a decrease in resistivity of the substrate. The fourth result of Fig. 17 show that distortion is mainly due to dispersion and attenuation in signal is more due to the conductor loss as compared to the dielectric loss even on low-resistivity substrate $\rho_s = 10 \Omega \text{ cm}$. Figure 18 further shows distortion in $\tau = 10$ ps square pulse dispersion and attenuation with respect to increasing distance for $w/h = 1.0$, $h = 0.45 \text{ mm}$, $t/h = 0.05$, $\epsilon_r = 11.9$, $\tan \delta = 0.001$, $\sigma_0 = 3.8 \times 10^7$ S/m and with respect to increasing strip width for $L = 3 \text{ mm}$. The distortion and attenuation significantly increase with an increase in distance. However, the increase in width has a strong influence. The narrow line is more dispersive causing more distortion and also on the low-loss substrate for higher conductor loss. The wide line has less distortion but more loss if the substrate is lossy.

Figure 19 shows variations in the rise-time and delay-time with respect to conductivities of six strip conductors for 50 ps pulse propagating on microstrip with $w = 0.3 \text{ mm}$,

Fig. 18 Distortion in 10 ps pulses with **a** $L=0, 3, 6, 9, 12$ mm for $w/h=1.0$ **b** $w=1, 10, 100, 300, 450, 4500$ μm for $L=3.0$ mm

$h=0.45$ mm, $w/h=0.67$, $\epsilon_r=11.9$, $t=0.0005$ mm, $L=3$ mm. Three substrate resistivity $\rho_s = \infty, 100, 10$ Ω cm are considered. Results are shown for both the direct and circuit

models. The rise-time is not influenced by the conductivity of strip conductor.

Likewise, it is not influenced by the substrate resistivity above $\rho_s > 100$ Ω cm. However, it is influenced by the substrate resistivity below. The rise-time is reduced from 10.5 to 6 ps with a decrease in resistivity from 100 to 10 Ω cm. The delay-time is reduced from 17 to 5.5 ps with a decrease in resistivity from 100 to 10 Ω cm. The low conductivity of the strip conductor has an influence on delay-time. For the high resistivity substrate, $\rho_s > 10$ Ω cm, the delay-time decreases to some extent below conductivity 3×10^7 S/m, from 17 to 16 ps.

However, for low-resistivity substrate 10 Ω cm, it increases to some extent, from 5.5 to 6.0 ps. These results are obtained using the circuit model. Figure 20 shows the effect of strip conductivity on overshoots and undershoots of 50 ps square pulse on microstrip with $w=0.3$ mm, $h=0.45$ mm, $w/h=0.67$, $\epsilon_r=11.9$, $t=0.0005$ mm, $L=3.0$ mm. Results are shown for substrate resistivity $\rho_s = 100$ Ω cm and $\rho_s = 10$ Ω cm. On a high resistivity substrate, both the direct and circuit models provide almost identical results. However, for low-resistivity substrate, $\rho_s = 10$ Ω cm, results differ significantly. More accurate circuit model shows significantly lower overshoot and undershoot. Low conductivity strip conductors have a lower amplitude of overshoot and undershoot. It is due to high conductor loss. However, we noted above that it also increases attenuation for the main body of the pulse. It is not desirable.

4 Conclusion

This article has been analyzed the propagation characteristics of transient pulses over the designed THz interconnects. The Fourier transformation approach has been used

Fig. 19 Effect of strip conductivity on **a** rise-time **b** delay-time on 50 ps pulse

Fig. 20 Effect of strip conductor conductivity on under-shoot and overshoot of 50 ps pulse

Table 6 Major advancements of the proposed work with the references [53–56]

Signal frequency	Interconnect materials	Description	Technology	References
50 GHz	Aerosol-jet silver	Optical interconnects	CMOS	[53]
10, 45 GHz etc.	Gold, poly-Si, SOI etc.	Optical + plasmonic interconnects	CMOS, BiCMOS, FPGA etc.	[54]
2.5 GHz	Copper	Hybrid (wireless + electrical interconnects	CMOS	[55]
0.5 THz	Multimode polymer	Optical interconnects	CMOS	[56]
10.0 THz simulated	Copper, gold, aluminum, tungsten, tungsten-silicide and poly-Si	Electrical interconnects	CMOS	[P]

P represents proposed work

to visualize the effect of propagation made by the various conducting materials which are widely used in CMOS technology and microwave integrated circuits as shown in Table 6. All parameters of transient pulse have been controlled by using integrated/circuit models and it has been verified with simulation results. The analyzed results could be useful in designing both on board and on-chip interconnects. All propagation effects are of importance in the performance analysis of designing integrated circuits interconnects in modern technologies.

Compliance with ethical standards

Conflict of interest The authors declare that they have no conflict of interest.

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