Transmission line characterization and modeling for electronic circuits and systems design

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ABSTRACT

Channel bandwidth-limited high-speed links or interfaces make circuit solutions not efficient. Both recent and subsequent links (SerDes-Serializer/Deserializer) design demand efficient and effective coupling between future circuit design, communication, and optimization. The challenges vary and new solutions are needed. In this article, an analytical wireline model is presented to predict electronic path loss towards adequate designs of electronic circuits and systems. An open loop system analysis is adapted in this paper. Our model was tested against different channels: a legacy channel with via stub discontinuity and FR4 dielectric, and a more recent microwave-engineered channel without stub and NELCO 6,000 dielectric, a very good matching attained. Good agreement was observed between our model and electromagnetic full-wave simulation data, as a result showed high level of applicability to thin-film microstrip line for adequate circuit design. The model is recommended for electronic engineers for adequate and faster interfaces and high-speed links designs.

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1. INTRODUCTION

High-speed links or interfaces (SerDes) are generally limited by channel's bandwidth which consequently affects the outputs of electronic communication links (such as system in package, and chip-to-chip) as signal has to propagate different traces/channels in order to get to its destination from source. This problem makes electronic channel characterization inevitable for adequate electronic circuits and systems designs. Over the channel, there is increase line attenuation or loss with frequency caused by dielectric loss and skin effect [1]. The line loss causes another low pass signal filtering, and more negative impacts from short traces (such as vias, connector, and traces) linking electronic components together. Many authors have attempted and proposed nomograms and channel/transmission line models for appropriate expected signal level prediction in chip-to-chip, system in package links [2]-[9]. Several works have been done on electrical simulation of interconnects with various techniques developed, such as resistance-capacitance (RC) tree interconnect representation, and two-pole approximation technique [10], [11]. However, they are based on approximations with underlying assumptions, the degree of accuracy is compromised while enhancing the speed. This makes them almost undesirable from an optimization view point, because the results of optimization depend on accuracy of the models applied. In [12]-[16] gave advanced technique which reduced the simulation time of interconnect during analysis but cannot give the simulation speed needed for adequate iterative optimization. Due to inaccessibility of suitable methods for on-line iterative optimization, it has given rise to some simple, fast on-line technique, such as polynomial curve-fitting methods in interconnect analysis were also used by [17], [18]. Nevertheless, curve-fit technique handles just mild nonlinearity with small parameters per time. Look-up methods have been employed to minimize simulation time of interconnects analysis [19]. However, they have many notable shortcomings as stated in [8].

Recently, low cost, light weight and small size features of microstrip antennas have made them desirable and appropriate for professional and commercial applications, and mobile wireless communication systems. Microstrip antenna is easier to integrate with electronics and adaptable to hybrid and monolithic integration circuit fabrication at microwave frequencies [20]-[22]. This has given rise to a reconfigurable switchable single and double notched band microstrip slot antenna [20], compact printed circuit board (PCB) antenna for use in microsatellites [23], and stacked microstrip antennas [22] among others. However, all transmission lines including microstrip lines suffer from multiple power loss mechanisms, such as dielectric loss, radiation loss and conductor loss, with dielectric loss and conductor loss being mostly dominant [24]. Therefore, microstrip trace or channel characterization and behavioral modeling become a necessary task for adequate designs. In this paper, a simple, fast and adequate wireline stochastic channel model is proposed for circuits and systems channel characterization in electronic system designing (high-speed links SerDes). The model is suitable for present and future links that requires effective coupling between high-speed systems. The rest of the paper is organized as follows: Section 2 describes the proposed model. Section 3 shows the results of simulation and the agreement between the model and baseline channel measurement. Section 4 is the application of our model to thin-film microstrip line (TFMSL) and Section 5 is the conclusion section of this article.

2. PROPOSED WIRELINE CHANNEL MODEL

A wireline is characterized as an open loop system having both the static and dynamic variations of the medium. Figure 1 shows a typical open loop system, x(t) is the input signal, l is the length of the wireline, h(t) is the response of the medium (wireline), n(t) is the noise and y(t)/l is the output received signal per length which is expressed as (1),

$$y(t)/l = h(t)x(t) + n(t) \tag{1}$$

a propagating signal along a wireline exhibits inverse proportionality to a particular degree m of its length; signal intensity per.

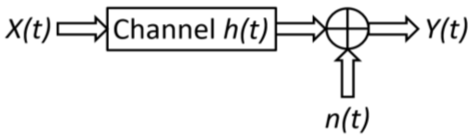

Figure 1. The system design

Length can generally be represented as (2) and (3),

$$\frac{y(t)}{x(t)} \propto \frac{\lambda}{4\pi l^m} \tag{2}$$

$$\frac{y(t)}{x(t)} = \frac{K\lambda}{4\pi l^m} \tag{3}$$

where λ is the wavelength of the signal and K is the constant of proportionality that defines links parameters and electrical resistivityρ, ϵ_e denotes the effective dielectric constant, and dielectric regions of the microstrip of the wireline material; which depends on the kind of material employed, generator gain G_t, receiving station gain G_m, K is therefore (4),

$$K = \epsilon_e \rho G_t G_m \tag{4}$$

where for a microstrip line [25].

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\frac{1}{\sqrt{1 + \frac{12d}{W}}}$$

W is the width of conductor, *d* is thickness of the grounded dielectric substrate, and ϵ_r is the relative permittivity. Also, a propagating signal in a medium varies with frequency, so we therefore introduce a factor gf^n. *f* is frequency, *n* and *g* account for other losses due to via-hole, and bends. of the wireline. Therefore (4) can be expressed as (5),

$$\frac{y/l}{x} = h = \frac{\epsilon_e \rho G_t G_r c}{4\pi\left(\frac{l}{l_0}\right)^m g\left(\frac{f}{f_0}\right)^n l_0^m f_0^n} \quad (5)$$

practically, there is no ideal environment, then $n > 0$. Assuming noise *n(t)=0,* then, at a given observation point, the received signal is (6) and (7).

$$y(t)/l = \frac{\epsilon_e \rho G_t G_r c}{4\pi\left(\frac{l}{l_0}\right)^m g\left(\frac{f}{f_0}\right)^n l_0^m f_0^n} x(t) \quad (6)$$

$$P_r = \frac{P_t G_t G_r}{\sigma} \quad (7)$$

Hence, in (5) can predict intensity of signal at a given point. However, loss σ and intensity of the received signal share inverse relationship between eachother (7) [2], using substitution and logarithmic technique, loss is expressed as (8) and (9):

$$\sigma(dB/l) = \sigma_0(l_0, f_0, \rho, d, W, \epsilon_r) + 10mlog\left(\frac{l}{l_0}\right) + 10nlog\left(\frac{f}{f_0}\right) + P \quad (8)$$

where

$$\sigma_0(l_0, f_0, \rho, d, W, \epsilon_r) = 10\log\left(\frac{4\pi l_0^m f_0^n}{\left(\frac{\epsilon_r+1}{2} + \frac{\epsilon_r-1}{2}\frac{1}{\sqrt{1+\frac{12d}{W}}}\right)\rho c}\right) \quad (9)$$

Frequency *f* is in *Hz*, the wireline length *l* may be measured in *mm*, *µm*, *cm*. depending on area of deployment, l_0 is the reference length, f_0 is in *Hz*, $\sigma_0(l_0, f_0, \rho, d, W, \epsilon_r)$ is the reference attenuation in *dB*, *c* is the speed of light (*m/s*), *P* is distortion due to via hole, and bends. which is frequency dependent. ρ is electrical resistivity of the wirelines such as copper wire, microstrip line, stripline, waveguide, and optical fiber. which is usually constant. Exponents: *m* and *n* can be determined by parameter extraction from measurement data using (10), where *N* is the number of points, E is the expression obtained from our attenuation model at different points. For instance, one data point equals the difference between the observed data and model data, carry out partial differential equation with respect to *m* and *n* and solve the consequent equations simultaneously.

$$S(n, m) = \sum_{i=1}^{N} E_i^2\,(n, m) \quad (10)$$

3. SIMULATIONS

The proposed model was simulated based on electrical resistivity of copper $\rho = 1.68 \times 10^{-8}$ at 20 ^{0}C with an assumption that relative permittivity $\epsilon_r = 1$ which in turn makes effective dielectric constant unity. The reference frequency was assumed 1 *GHz,* reference length $l_0 = 1\ m$, from (8); *P=0* (without distortion), wireline length $l = 0.66\,m$ and frequencies ranges between 5 and 40 GHz. Estimated reference loss $\sigma_0 \approx 76\ dB$ ($at\ m = 1, n = 0.8$) using (9). This reference loss value may not be appropriate for all materials but *P* value/function will always compensate for the inappropriateness. Simulation was performed for *n=0.8, 0.9, 1* at *m=1*, however, *m* and *n* values are statistical values obtainable from measurement data. Results obtained are as shown in Figure 2. The model shows a high degree of agreement with various path loss models; path loss increases as transmission frequency increases for a specific material length. Change in correction factors, which account for other losses leads to change in path loss along the wireline.

Determining the Shannon capacity for a medium with practical noise sources is important, because: it estimates the usable channel bandwidth (i.e. the requirement for circuit speed); and it determines the possible speed of data. These factors become crucial in predicting the lifespan of copper wires as a signaling channel. Our model's analysis (test) and validation were based on two channels, shown in Figure 3 of [3] because it represents the two opposite sides of all the range of media over which present SerDes are needed for operation. The first medium denotes a class of older media with impedance discontinuity originated from via stubs and connectors, leading to frequency domain notches, and older dielectrics, FR4, and leading to a higher loss. The second medium was designed for reduction in the impedance discontinuities from connectors and via stubs, and as a result has plane roll-off. It equally exhibits smaller loss slope in NELCO6000 dielectric.

Our model was applied with the same conditions such as wire length *l=26 inches (0.66 m)*, reference length and reference frequency remain unchanged. Parameter *m* was determined to be *1* for all scenarios and *n* was equally determined to be *0.8* and *0.87* for NELCO (no stub) and FR4 (via stub) respectively. Result obtained in Figure 3 show a high level of agreement of our model with the baseline channel.

Figure 4 shows different values of *P* obtained at different frequencies for the two materials under test and *P* function was modeled out (11). The parameter coefficients curve fitted in Figure 5 are summarized in Table 1. Hence, our model is a good channel behavior prediction tool for circuits and systems design,

$$P = \alpha_o + \alpha_1 \log\left(\frac{f}{f_o}\right) + \alpha_2 \left(\log\left(\frac{f}{f_o}\right)\right)^2 \quad (11)$$

we also used the NELCO (no stub) material condition as a slave model to test our model's performance under various conditions.

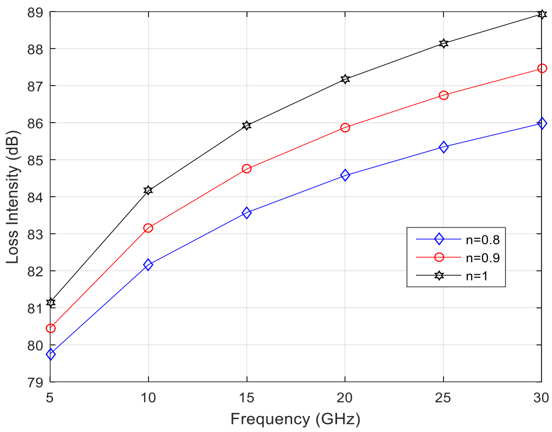

Figure 2. Estimated loss intensity versus frequency for the correction factor scenarios

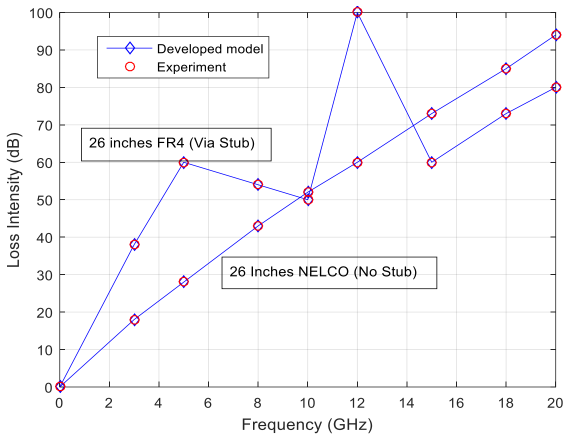

Figure 3. Model validation against baseline channels

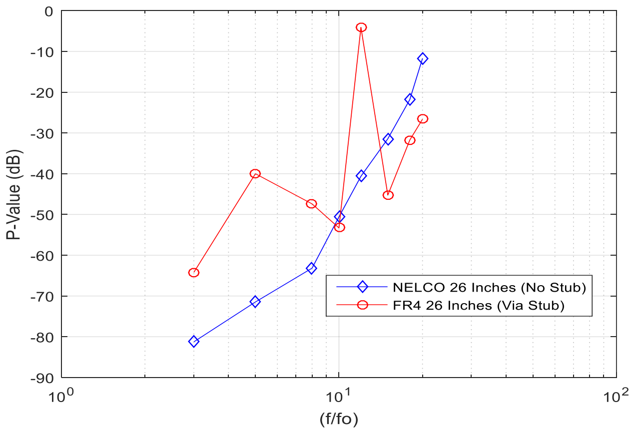

Figure 4. P-value against frequency

Table 1. Estimated parameters of the model

Material	α_o	α_1	α_2	m	n
NELCO (no stub)	*-93.38*	*4.25*	*-0.01*	1	0.8
FR4 (via stub)	*-86.27*	*7.55*	*-0.24*	1	0.87

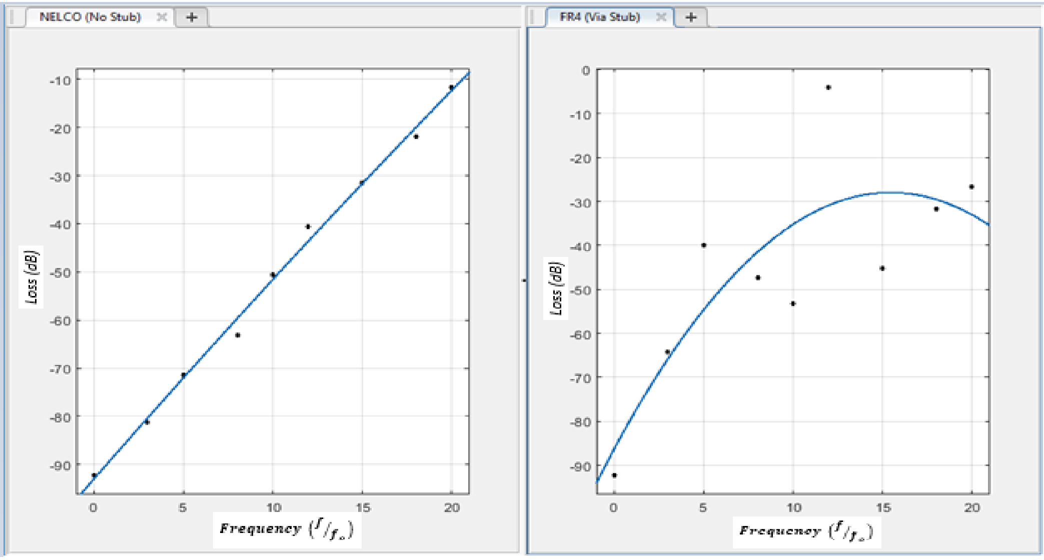

Figure 5. P-coefficients curve fitting results for the materials NELCO—no stub and FR4—via stub materials

Relative permittivity of different materials; alumina (99.5%), beeswax, beryllia and glazed (all at *10 GHz*) were tested with *d=0.01 mm*, and *W=0.03 mm*, result obtained is as shown in Figure 6, loss decreases with increase in material relative permittivity. Also, varying the thickness of the wireline for alumina 99.5% material and width *W=0.03 mm*, Figure 7 shows the simulated result; loss increases with material thickness of wireline.

Another condition considered is varying width *W* of alumina 99.5% material with thickness *d=0.01 mm*. Result obtained is as shown in Figure 8. Loss intensity decreases as width *W* of the wireline increases. Finally, different materials electrical resistivity at 20 ^{0}C: copper, aluminium, silver and tungsten were examined with unity relative permittivity (which makes the effective relative permittivity constant unity). Figure 9 shows the model's prediction at different materials considered; loss intensity reduces with increasing wireline material resistivity value.

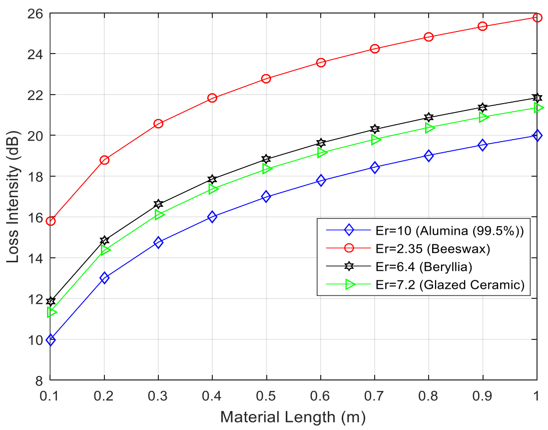

Figure 6. Model loss prediction with various relative permittivities at 10 GHz and 20 ^{0}C (*Tangential factor: Beewax=0.005, Beryllia=0.0003, Glazed ceramic=0.008, Alumina (99.5%) =0.0003*)

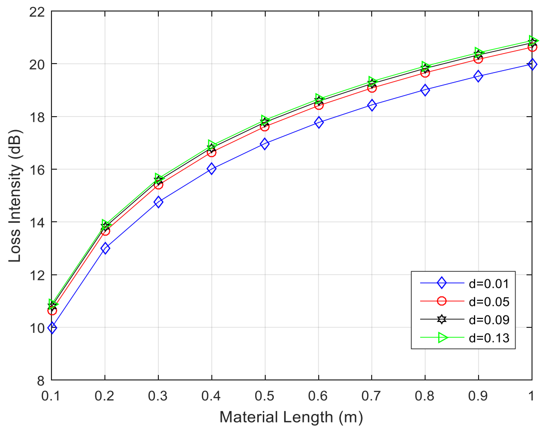

Figure 7. Model prediction along wireline with different thickness level

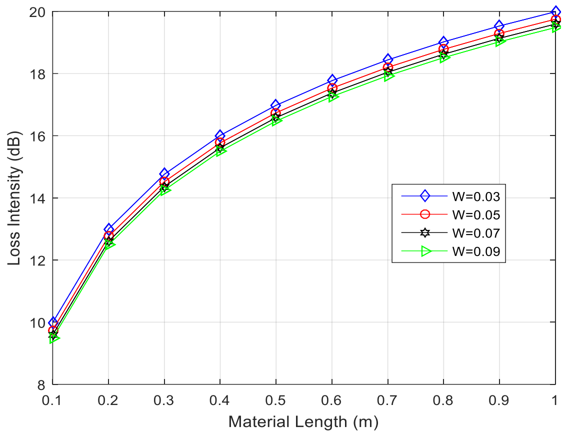

Figure 8. Model loss prediction along a wireline with various material widths

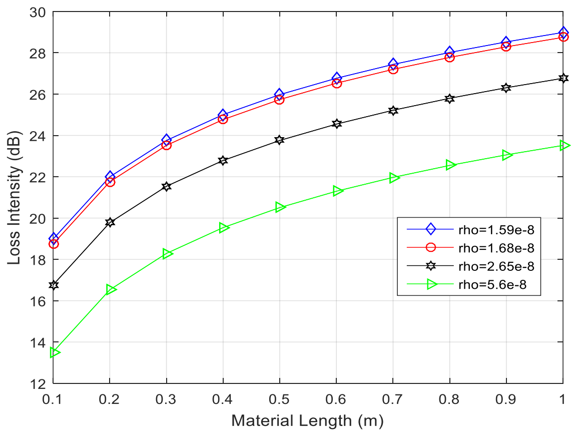

Figure 9. Loss prediction through wireline at different resistivity values

4. THIN-FILM MICROSTRIP LINE

TFMSLs are specially used in silicon-based monolithic microwave integrated circuits (MMICs) and used as transmission line in multichip module (MCM) [26]. Availability of high-quality polymers such as Benzocyclobutene (BCB), and polymide, makes losses in TFMSLs likened to coplanar waveguides (CPWs) in GaAs MMICs. In addition, because of the scalability of transversal dimension, reduced TFMSLs exhibits great little-dispersive features, and it's useable at sub-millimeter-wave range. Differences in geometrical dimensions: width, height, thickness and conductivity of metallization cause a degree of discrepancies in the electrical response of signal conductor. Our model was tested against a robust TFMSL $(BCB\ with\ \epsilon_r = 2.7\ and\ tan\delta_z = 0.015;\ W = 8\ \mu m, h_s = 1.7\ \mu m, d = 0.8\ \mu m, Wg = 8\ \mu m;$ conductivity of metalization $k = 2.5 \times 10^7 S/m)$ full wave EM simulation data in Figure 6 of [26]; result obtained is as shown in Figure 10. P-values against the corresponding frequencies behavior and curve fitting response of P-function are as shown in Figure 11 respectively. The model parameters are as summarized in Table 2. With this condition, our model can replace the complex one in [26]. This model is derived from stochastic point of view and effortlessly administered in different softwares. It is appropriate for conventional microstrip structure as well. The frequency range of application for TFMSLs can be extended down to dc level and upper limit reaches submillimeter-wave range. The model is a good tool for channel or interconnecting loss/attenuation estimation for electronic circuit and system engineers for effective design.

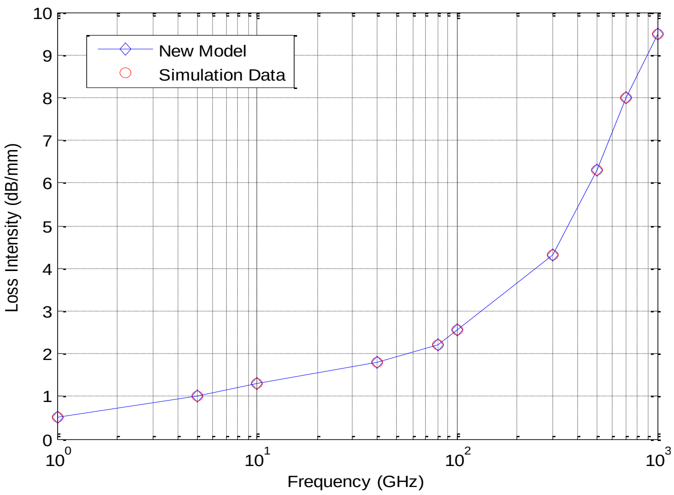

Figure 10. Model against EM simulation data

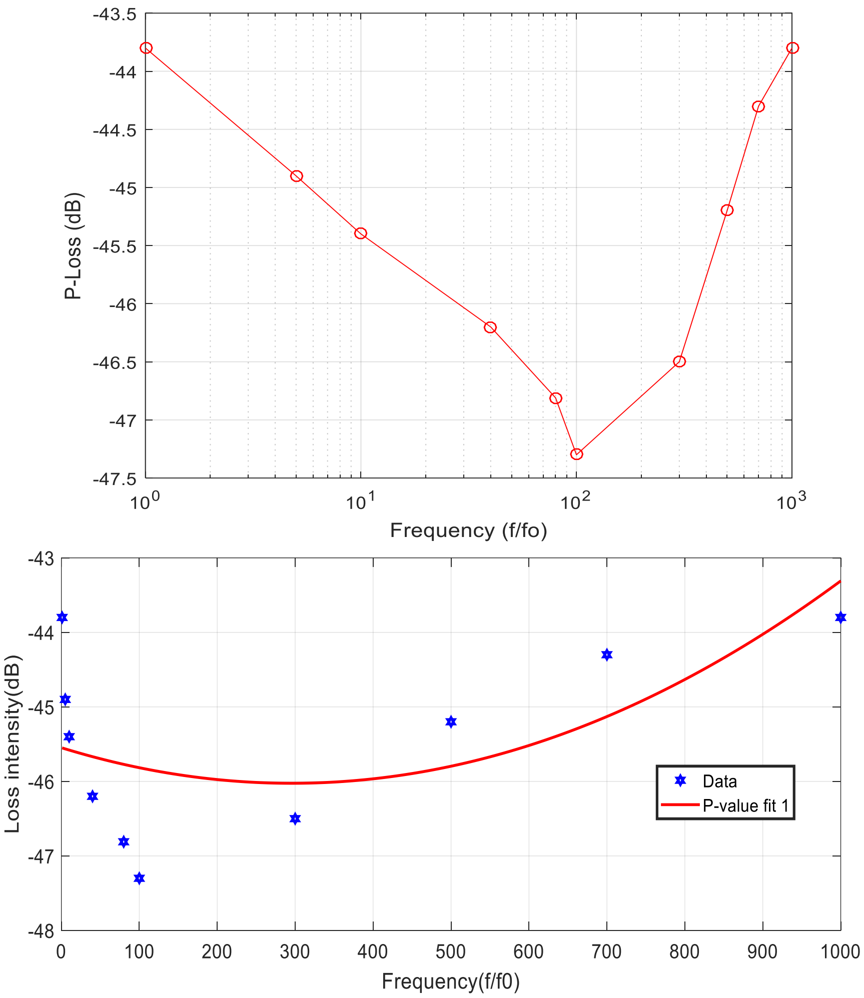

Figure 11. P-values against frequencies and curve fitting response

Table 2. Coefficients parameter

Parameter	m	n	α_o	α_1	α_2
Value	*1*	*0.3*	*-45.55*	*-0.003*	*5.5E-6*

A 2-port network S-parameter data for a lossy transmission line (12) [4], [25], and [26] were considered against the developed model in terms of speed (simulation time) using MATLAB at different frequencies:

$$\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} \frac{(z_{out}^2 - z_0^2)\sinh\gamma l}{(z_{out}^2 - z_0^2)\sinh\gamma l + 2z_0 z_{out}\cosh\gamma l} & \frac{2z_0 z_{out}}{(z_{out}^2 - z_0^2)\sinh\gamma l + 2z_0 z_{out}\cosh\gamma l} \\ \frac{2z_0 z_{out}}{(z_{out}^2 - z_0^2)\sinh\gamma l + 2z_0 z_{out}\cosh\gamma l} & \frac{(z_{out}^2 - z_0^2)\sinh\gamma l}{(z_{out}^2 - z_0^2)\sinh\gamma l + 2z_0 z_{out}\cosh\gamma l} \end{bmatrix} \quad (12)$$

where Z_{out} is the characteristic impedance of the transmission line, Z_o is the port impedance of the network analyzer, $\gamma = \frac{2\pi f}{c}$ (for a generalized lumped-element model of a transmission line) is propagation constant of the line and l is length of the line. Since transmission lines are symmetrical and reciprocal networks $S_{11} = S_{22}$ and$S_{12} = S_{21}$. Using an HP COMPAC laptop with 4 GHz RAM, simulation time test result obtained is as summarized in Figure 12. It is evident that the developed model is faster; hence channel can now be substituted by an empirical equation as against S-parameter data set out of EM-simulation tool which helps to speed up simulation/design of circuit or system.

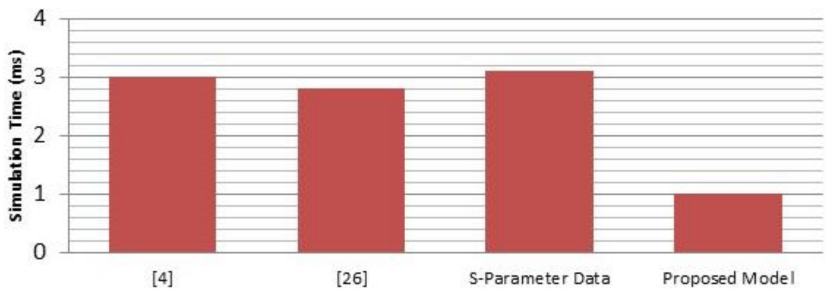

Figure 12. Speed test result

5. CONCLUSION

This paper proposed a new and faster signal loss/attenuation model for SerDes wireline channel modeling or high-speed links for electronic circuits and systems design. In this article, an analytical wireline model is presented to predict electronic path loss towards adequate designs of electronic circuits and systems. An open loop system analysis is adapted in this paper. Our model was tested against different channels: a legacy channel with via stub discontinuity and FR4 dielectric, and a more recent microwave-engineered channel without stub and NELCO 6,000 dielectric, a very good matching attained. Good agreement was observed between our model and electromagnetic full-wave simulation data; as a result, showed high level of applicability to thin-film microstrip line for adequate circuit design. The resulting model shows high level of agreement with full wave EM simulation data and therefore applicable to thin-film microstrip line interconnecting/coupling. This model is simple, faster and suitable for adequate circuits and systems link design.

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