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Minimizing crosstalk for high-speed and high-density bus systems using the sample-decision method①

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Abstract

This paper presents a method based on a sample-decision (SD) circuit to suppress crosstalk and noise for a high-speed and high-density bus system. A method to count the number of times of SD for different length of transmission lines is presented and a bit error rates (BERs) formula is given by the SD circuit. It is shown that for long transmission line systems, multiple SD circuits can improve the BERs significantly. Circuits simulation for single SD method is also done, it is found that when the amplitude peak values of the superposed crosstalk and noise are less than half of the corresponding signal ones, they will be eliminated completely for the cases investigated.

Key words: crosstalk, transmission line, sample-decision, high-speed and high-density circuits

0 Introduction

In a high-speed, high-density bus system, electromagnetic wave coupling between closely spaced signal lines limits interconnect performance[1]; thus, minimizing the crosstalk among integrated signal transmission lines, such as microstrip lines, has become an important goal in the design of high-speed, high-density circuits. Crosstalk, which is the transfer of an unwanted signal from one trace or circuit to its adjacent ones, is one of the major challenges in today's signal integrity for high-speed and high-density circuits. In practice, the crosstalk distorts communication signals and deteriorates the quality of digital and analog signals. It is found that crosstalk exists everywhere inside an integrated circuit, such as inside chips, printed circuit boards (PCB), interconnect packages, or any non-shielded high density integrated circuits[2,3]. Consequently, investigation for crosstalk minimization is an important task in the design of high-speed, high density circuits.

Crosstalk takes place mainly due to the mutual coupling, e. g., mutual capacitance and mutual inductance, for any arbitrary two adjacent transmission lines. These lines are usually called victim and aggressor lines, respectively. When a signal passes through the aggressor line, crosstalk signals on the victim lines are introduced, since the coupling from the mutual capacitance and mutual inductance produces the energy transferred from the aggressor lines to the victim ones[4]. There are two major signal factors that greatly affect crosstalk: the instant voltage transition, namely, low to high or high to low[5], and mutual inductance and capacitance. When a transmission line delivers high frequency digital signals with relatively short rise and fall time, the corresponding transient voltage conversion will certainly yield serious crosstalk towards its adjacent traces. Moreover, it is found that a high-density circuit leads to more mutual inductance and capacitance crosstalk.

A common practice to reduce crosstalk is to increase spacing between traces, or to add an additional guard trace for separation of trace signals, or to change the shape of the traces[2-4,6-9]. Moreover, it was suggested that modal voltages and currents were applied for reduction of crosstalk and echo, which required particular circuits at sources and loads to drive and receive signals[10]. In addition, it was achieved by intentional skewing of relative timing of adjacent wires[11]. Also, it was found that the coding methods can be applied for reduction of crosstalk as discussed in Refs[12-14]. However, these methods for reducing crosstalk did not care about other noises. Different

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from the previous methods for minimizing crosstalk, we present a new crosstalk suppression scheme in this paper based on a digital communication principle——the sample-decision (SD) method. The SD method is commonly used in a digital baseband communication system. In the scheme, receivers' signals are regenerated by sampling and decision. In this research, a simple digital baseband transmission system is used to suppress the crosstalk and noises, and it is shown that this scheme can effectively reduce crosstalk and noise for the cases investigated.

1 Principle of the SD method

In a typical binary digital base-band communication system, if $g(t)$ is the sending signal, the receiving signal is

$$r(t) = s(t) + n(t) \tag{1}$$

where $s(t)$ is a function of the sending signal $g(t)$, channel characteristics and the receiving filter, and $n(t)$ is the Gaussian white noise. In a digital baseband communication system, the SD technique is applied at the receiver end in order to rebuild transmitted signal $g(t)$.

In a similar way, the receiving signal on the victim line in a bus system is given as

$$r(t) = g(t) + n(t) + c(t) \tag{2}$$

where $g(t)$ is the transmitted signal, $n(t)$ is Gaussian white noise, and $c(t)$ is the induced crosstalk.

Comparing Eqs (1) and (2), their representations are very similar except the additional unwanted crosstalk term $c(t)$. The method adopted in a digital baseband communication system can be directly applied to eliminate crosstalk and noise. The system model presented in this paper is shown in Fig. 1.

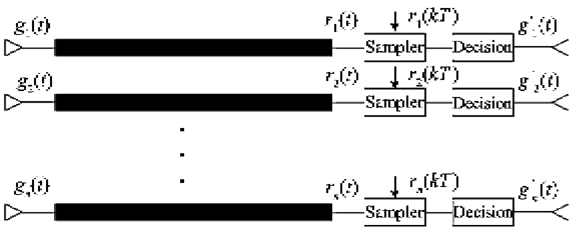


Fig. 1 Minimizing crosstalk model based on the sample-decision method

In this model, it is assumed that the transmitted signal is the unipolar NRZ binary code, and that each bit of the code amplitude is 0 or A, representing digital signal "0" or "1", respectively. A digital transmission signal $g(t)$ with the time period of T is given as

$$g(t) = \begin{cases} 0, & \text{data "0"} \\ A, & \text{data "1"} \end{cases} \tag{3}$$

If the power loss is negligible in signal transmission, then at the discrete sampling time $t_k = kT$, the amplitude of $r(t_k)$ is

$$r(t_k) = n(kT) + c(kT) \tag{4}$$

or

$$r(t_k) = A + n(kT) + c(kT) \tag{5}$$

By applying the SD method in the transmission line model, at its receiver, we set the decision threshold to be $v_d = A/2$. The decision rule is that if $r(t_k) < v_d$, it is set to be "0"; otherwise, to be "1". This means that the digital signals distorted by crosstalk and noise can be rebuilt correctly with a simple sample and decision circuit, provided that the amplitude of crosstalk and noise is less than the threshold voltage.

2 Single and mutiple section SD circuits

The proposed method herein is to rebuild the originally transmitted signals by using a sample-decision circuit at the receiver end. Since the sample clocking of the circuit usually takes place at the middle of digital signal pulses, the crosstalk for the regenerated digital signals can be significantly suppressed after the SD process. It is found that crosstalk increases greatly as the length of transmission lines augments, which implies that a single section SD circuit may not be enough to eliminate the crosstalk introduced symbol errors; in other words, multiple sections of the SD circuits may be needed to completely resolve the issue. An immediate question is how to determine the number of the SD circuit sections. This will be addressed in the following discussion.

In a high speed and high density bus system, it is assumed that the amplitude of digital signals on the adjacent aggressor line is v_a, rising time of signals is RT, and the length of the lines is l. The maximum peak value of crosstalk on the far-end of victim line is given as Ref. [3]

$$v_f = \frac{l}{\mathrm{RT}} \times k_f \times v_a \tag{6}$$

and

$$k_f = \frac{1}{2v} \times \left(\frac{C_m}{C_L} - \frac{L_m}{L_L}\right) \tag{7}$$

where v is the speed of the signal on the line, C_m and C_L are the mutual capacitance and capacitance per unit length of the trace, and L_m and L_L indicate the mutual inductance and inductance per unit length of the trace.

In order to implement the previous developed SD techniques for accurately receiving signals, the threshold of decision v_d is setup to be $v_a/2$, where v_a is the magnitude of signal voltage. For simplicity, we further assume that crosstalk is far greater than other noises, which can be, thus, ignored in the analysis. When the peak value of crosstalk reaches the decision threshold, namely $2v_f = v_a/2$ by considering that the crosstalk on a victim line is mainly introduced from its two nearest aggressor lines[3], we define an effective SD length of the transmission line as the length needing a SD circuit section denoted by l_f. According to Eq. (8), we have

$$l_f = \frac{\mathrm{RT}}{4k_f} \tag{8}$$

From the calculated l_f and the physical trace length l of the transmission lines, we can determine the number of the SD circuits, N, which is discussed in the following two cases:

Single SD Circuits, $l < l_f$:

In this case, $v_f < v_a/4$, only a single SD circuit ($N = 1$) is needed. Because the crosstalk will not be able to reach the threshold of decision, there is no symbol error after sample-decision, and the system SD model is demonstrated in Fig. 1.

Multiple SD Circuits, $l \geqslant l_f$:

In this case, however, multiple section SD circuits have to be used to minimize crosstalk. l can be divided into $l = nl_f + l_a$, where $n = 1, 2, 3, \cdots$, l_f is defined in Eq. (8), and l_a is a transmission line length with $l_a < l_f$. Each l_f section employs one SD circuit, plus the last SD circuit used by the remaining line section l_a. Obviously, the number of the SD circuit sections is $N = n + 1$, and the associated system SD model is displayed in Fig. 2.

3 Crosstalk effect on bit error rate

As a consequence of digital signal transmission along a trace, the crosstalk towards its adjacent lines can eventually reduce the bit error rates for digital communication. In the following, based on the proposed sample-decision method, we estimate the bit error rate of the digital transmission for the cases investigated. It is known that the probabilities appearing "0" and "1" at any arbitrary sampling point are identical and that any two continuous bits, namely, "10", "01", "00", and "11" have the equal possibility of 1/4. Since signals "00" and "11" don't involve a signal conversion, no crosstalk is induced in these circumstances. Thus, the crosstalk and its corresponding appearing probabilities would be discussed in three cases as follows:

- Case Ⅰ. Signals changing from "1" to "0", crosstalk would be v_f and its probability of appearance is 1/4;
- Case Ⅱ. Signals changing from "0" to "1", crosstalk would be $-v_f$ and its probability of appearance would be 1/4;
- Case Ⅲ. No changing, crosstalk would be 0 and its probability of appearance would be 1/2.

Since crosstalk signals can be added repeatedly on the victim line in the case of two adjacent aggressor lines, the peak values could be $2v_f$, $-2v_f$, v_f, $-v_f$ and 0. According to the probability theory[15,16], the probabilities of appearance would be: 1/16, 1/16, 1/4, 1/4, 3/8 respectively, and other noises are usually treated as the Gaussian white noise, and its BER can be evaluated accordingly.

As seen in Appendix, it can be shown that when $l < l_f$, the BER of the single SD circuit on receiver for the worst case (sampling moment point to the peak of crosstalk) is shown as

$$\begin{aligned} P_b = {} & \frac{1}{32}\mathrm{erfc}\left(\frac{v_a + 4v_f}{2\sqrt{2\sigma^2}}\right) + \frac{1}{32}\mathrm{erfc}\left(\frac{v_a - 4v_f}{2\sqrt{2\sigma^2}}\right) \\ & + \frac{1}{8}\mathrm{erfc}\left(\frac{v_a + 2v_f}{2\sqrt{2\sigma^2}}\right) + \frac{1}{8}\mathrm{erfc}\left(\frac{v_a - 2v_f}{2\sqrt{2\sigma^2}}\right) \\ & + \frac{3}{16}\mathrm{erfc}\left(\frac{v_a}{2\sqrt{2\sigma^2}}\right) \end{aligned} \tag{9}$$

where P_b is the bit error rate, σ^2 is the power summarization of Gaussian noises except the crosstalk, v_a is amplitude of transmitted signal, and erfc (x) is the complementary error function which is defined in Ref. [15].

When $l \geqslant l_f$, the BER for the multiple section SD circuits is further studied, and its final expression is shown in Eq. (10). Here, we set $l_a = l_f/2$ as an example, then $l = nl_f + \frac{l_f}{2}$, and the number of the SD circuit sections is $N = n + 1$, which means we can determine the BER from the ratio (n) of length l and l_f, as given in Eq. (10). Table 1 shows the BERs for single and multiple SD sections when the *SNR* (signal to noise ratio) is 10 and 20dB, respectively. It is observed that the BER for a multiple SD circuit is always better than that of a single one, which are, especially more significant when *SNR* = 20dB for the cases investigated.

Table 1 Comparison of BER performance between the two methods

$l = nl_f + \frac{l_f}{2}$		*SNR* = 10dB	*SNR* = 20dB
$n = 1$, $l = \frac{3l_f}{2}$	$N = 2$	0.0926	0.0512
	$N = 1$	0.1310	0.1263
$n = 2$, $l = \frac{5l_f}{2}$	$N = 3$	0.1553	0.0809
	$N = 1$	0.2452	0.2405
$n = 3$, $l = \frac{7l_f}{2}$	$N = 4$	0.2136	0.1097
	$N = 1$	0.3056	0.3008

$$
\begin{aligned}
P_b = {} & 1 - \{1 - [\frac{1}{32}\mathrm{erfc}(\frac{v_a + 2\frac{v_a}{2}}{2\sqrt{2\sigma^2}}) + \frac{1}{32}\mathrm{erfc}(\frac{v_a - 2\frac{v_a}{2}}{2\sqrt{2\sigma^2}}) + \frac{1}{8}\mathrm{erfc}(\frac{v_a + \frac{v_a}{2}}{2\sqrt{2\sigma^2}}) + \frac{1}{8}\mathrm{erfc}(\frac{v_a - \frac{v_a}{2}}{2\sqrt{2\sigma^2}}) \\
& + \frac{3}{16}\mathrm{erfc}(\frac{v_a}{2\sqrt{2\sigma^2}})]\}^n \times \{1 - \{\frac{1}{32}\mathrm{erfc}[\frac{v_a + 2(2v_f - n\frac{v_a}{2})}{2\sqrt{2\sigma^2}}] + \frac{1}{32}\mathrm{erfc}[\frac{v_a - 2(2v_f - n\frac{v_a}{2})}{2\sqrt{2\sigma^2}}] \\
& + \frac{1}{8}\mathrm{erfc}(\frac{v_a + 2v_f - n\frac{v_a}{2}}{2\sqrt{2\sigma^2}}) + \frac{1}{8}\mathrm{erfc}(\frac{v_a - 2v_f - n\frac{v_a}{2}}{2\sqrt{2\sigma^2}}) + \frac{3}{16}\mathrm{erfc}(\frac{v_a}{2\sqrt{2\sigma^2}})\}\} \\
= {} & 1 - \{1 - [\frac{1}{32}\mathrm{erfc}(\frac{v_a}{\sqrt{2\sigma^2}}) + \frac{1}{32}\mathrm{erfc}(0) + \frac{1}{8}\mathrm{erfc}(\frac{3v_a}{4\sqrt{2\sigma^2}}) + \frac{1}{8}\mathrm{erfc}(\frac{v_a}{4\sqrt{2\sigma^2}}) \\
& + \frac{3}{16}\mathrm{erfc}(\frac{v_a}{2\sqrt{2\sigma^2}})]\}^n \times \{1 - [\frac{1}{32}\mathrm{erfc}(\frac{v_a - nv_a + 4v_f}{2\sqrt{2\sigma^2}}) + \frac{1}{32}\mathrm{erfc}(\frac{v_a - nv_a + 4v_f}{2\sqrt{2\sigma^2}}) \\
& + \frac{1}{8}\mathrm{erfc}(\frac{2v_a - nv_a + 4v_f}{4\sqrt{2\sigma^2}}) + \frac{1}{8}\mathrm{erfc}(\frac{2v_a + nv_a - 4v_f}{4\sqrt{2\sigma^2}}) + \frac{3}{16}\mathrm{erfc}(\frac{v_a}{2\sqrt{2\sigma^2}})]\}, \; n = 1,2,3,\cdots \quad (10)
\end{aligned}
$$

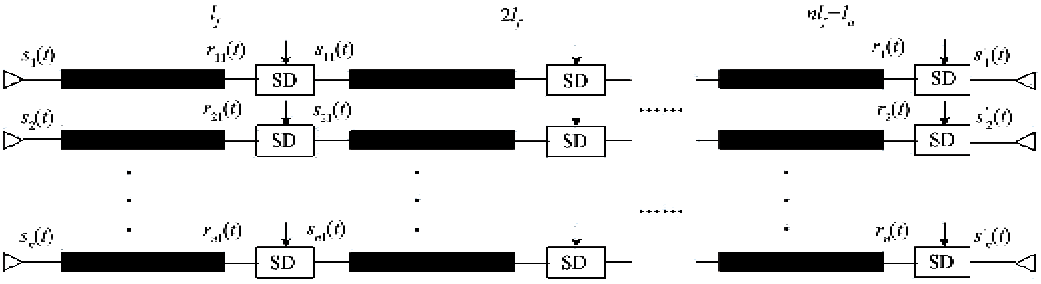


Fig. 2 Model for minimizing crosstalk based on the multiple sample-decision method

4 Verification of Crosstalk Elimination Method

This section is going to verify and demonstrate the the proposed SD method in minimizing crosstalk by using the advanced design system (ADS) software[17]. The cross-section of the high density signal traces (microstrip lines) is shown in Fig. 3 with specified dimensions and the length of the transmission line are 5 in ($5\text{in} < l_f$). The sample-decision circuit is placed at the end of transmission lines. The SD model for extracting time domain signals is developed as shown in Fig. 4. The transmitted and received signals are defined in Table 2:

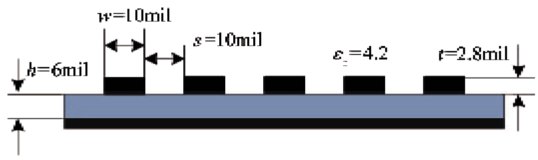


Fig. 3 PCB stack-up and view dimensions

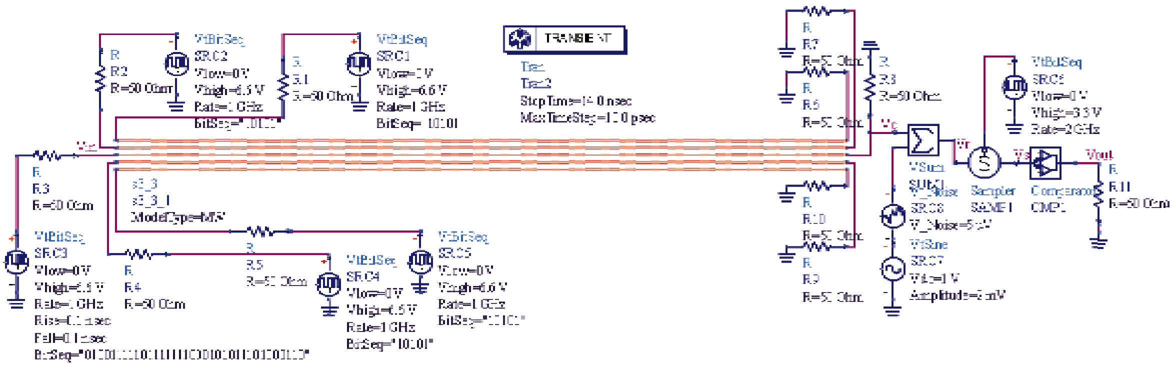
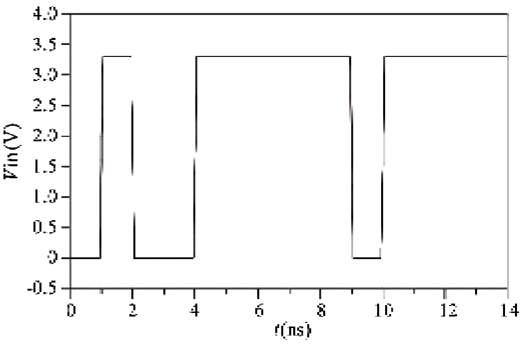


Fig. 4 Illustration of extracting the time domain signal

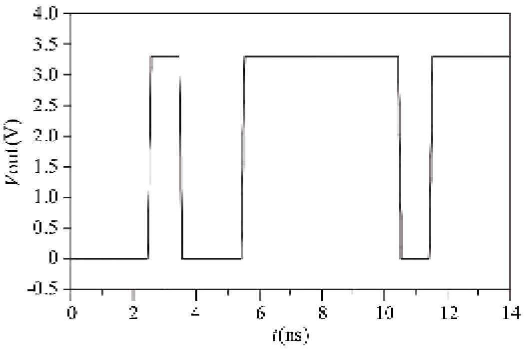
Table 2 The definition of the signal

Signal	Description
Vin	Input digital signal for transmission
Vc	Signal received at the end of the transmission line, which is affected by crosstalk
Vr	Vc added with noise
Vs	Vr sampled output
Vout	Deciding output signal

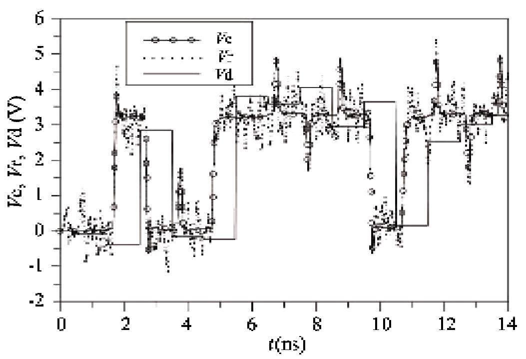
The input to the third signal trace is a pseudo-random sequence (unipolar binary code) with the period of 15ns, the code period of 1ns, and the amplitude of 3.3V. Input signals to other traces are 0 and 1 alternate inversion codes with the same code period and amplitude as those in the pseudo-random sequence. The decision threshold v_d is set to be 1.65V. Experimental results are plotted in Fig. 5. It is seen that that the



(a) Vin: Input digital signal for transmission



(b) Vout: Deciding output signal



(c) Signal: Vc, Vr, Vs

Fig. 5 Waveforms of various points in Fig. 4

crosstalk and noise have seriously distorted the digital signals. With the sample-decision circuits, the rebuilt signals are almost the same as the original ones, which indicates that the proposed SD method is effective in eliminating the unwanted crosstalk and noise.

Figs. 5(a) and 5(b) show the input and the SD output signals, which are essentially the same waveforms except that the later has a time delay, which is introduced by the propagation delay and the sample-hold delay. Notice that in the simulations sample clock is used. Usually, in a computer bus system or inside a chip, the sample clocks are provided by the main clock, which means the clock realization is much simpler than that in a digital baseband system because the sample clock in the sample-decision circuit does not need to be featured with bit synchronization.

5 Conclusions

This paper presents an efficient crosstalk elimination method by the application of the digital communication principle——sample-decision method. It is suggested that the number of the SD circuit sections is determined by evaluating the presented effective SD length l_f. Since the sample-decision method can make the sample clocking accurately point to the middle of digital signal pulses, it can maximize the signal to noise ratio. ADS simulation experiments show that a seriously distorted digital signal can be rebuilt successfully to efficient SD circuit when $l < l_f$. When a long line ($l \geqslant l_f$) is used, multiple SD circuits are suggested to improve the BER, which is deteriorated by crosstalk and other noises.

Appendix crosstalk BER derivation for single SD

Assuming that the receiving signal on the victim line is given as

$$r = a + n + v_c \tag{A-1}$$

where a is the transmitted signal, the amplitude is v_a or 0, n is Gaussian white noise, and v_c is the induced crosstalk, its peak values could be $2v_f$, $-2v_f$, v_f, $-v_f$, 0. Thus, the BER on the victim line can be evaluated accordingly as

$$P_b = P(e \mid a = v_a)P(a = v_a) + P(e \mid a = 0)P(a = 0) \tag{A-2}$$

Assuming the transmitted signal such as "0" or "1" is equal in probability, the BER is given as

$$P_b = \frac{P(e \mid a = v_a) + P(e \mid a = 0)}{2} \tag{A-3}$$

where $P(e \mid a = v_a)$ is the error probability of sending a signal "0", which is actually sending signal "1". It can be evaluated by using (A-4). Similarly, $P(e \mid a = 0)$ is the error probability of sending signal "1", in which signal "0" is transmitted instead. They can be evaluated as follows:

$$\begin{aligned} P(e \mid a = v_a) &= P(v_a + n + v_c < \frac{v_a}{2}) \\ &= P(n < -\frac{v_a}{2} - v_c) = P(n < -\frac{v_a}{2} - 2v_f)P(v_c = 2v_f) \\ &+ P(n < -\frac{v_a}{2} + 2v_f)P(v_c = -2v_f) \\ &+ P(n < -\frac{v_a}{2} - v_f)P(v_c = v_f) \\ &+ P(n < -\frac{v_a}{2} + v_f)P(v_c = -v_f) \\ &+ P(n < -\frac{v_a}{2})P(v_c = 0) \end{aligned} \quad (A\text{-}4)$$

$$\begin{aligned} P(e \mid a = 0) &= P(v_a + n + v_c > \frac{v_a}{2}) \\ &= P(n > -\frac{v_a}{2} - v_c) = P(n > -\frac{v_a}{2} - 2v_f)P(v_c = 2v_f) \\ &+ P(n > -\frac{v_a}{2} + 2v_f)P(v_c = -2v_f) \\ &+ P(n > -\frac{v_a}{2} - v_f)P(v_c = v_f) \\ &+ P(n > -\frac{v_a}{2} + v_f)P(v_c = -v_f) \\ &+ P(n > -\frac{v_a}{2})P(v_c = 0) \end{aligned} \quad (A\text{-}5)$$

Amplitude probability density function of Gaussian white noise is given as Ref. [15]

$$p(n) = \frac{1}{\sqrt{2\pi\sigma^2}} e^{\frac{-n^2}{(2\sigma^2)}} \quad (A\text{-}6)$$

where σ^2 is the power of Gaussian white noise.

Substitution of the relations (A-4) through (A-6) into (A-3), leads to the BER expression on the victim line for a single SD circuit.

$$\begin{aligned} P_b &= \frac{1}{32}\mathrm{erfc}(\frac{v_a + 4v_f}{2\sqrt{2\sigma^2}}) + \frac{1}{32}\mathrm{erfc}(\frac{v_a - 4v_f}{2\sqrt{2\sigma^2}}) \\ &+ \frac{1}{8}\mathrm{erfc}(\frac{v_a + 2v_f}{2\sqrt{2\sigma^2}}) + \frac{1}{8}\mathrm{erfc}(\frac{v_a - 2v_f}{2\sqrt{2\sigma^2}}) \\ &+ \frac{3}{16}\mathrm{erfc}(\frac{v_a}{2\sqrt{2\sigma^2}}) \end{aligned} \quad (A\text{-}7)$$

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