Asymmetric, Multi-Conductor Low-Coupling Structures for High-Speed, High-Density Digital Interconnects

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***Abstract*—Small inter-line spacings and high switching speeds emphasize the problems of crosstalk and coupling distortion in high-speed, high-density digital interconnects. The use of substrate compensation allows the design of structures where crosstalk and coupling can be essentially eliminated, even for inter-line spacings of less than one center conductor width. Some of the characteristics of this novel method are presented for both symmetric and asymmetric multi-line geometries. Pulse distortion and crosstalk on a four-line, symmetric structure is analyzed, showing how crosstalk and coupling distortion is reduced by substrate compensation. Pulse distortion on symmetric coupled lines is also studied, showing that it is possible to choose a substrate combination which significantly reduces coupling and crosstalk for a wide range of conductor configurations.**

Introduction

As CLOCK rates increase and inter-line spacings decrease, an accurate analysis of pulse propagation in multi-conductor, coupled microstrips becomes more important. Clock rates in the Gbit/s range with rise times on the order of 10 ps are being considered, resulting in pulses that contain significant frequency components well above 10 GHz. The small inter-line spacing used in high-density VLSI interconnects increases coupling and crosstalk, degrading the intended signal and introducing spurious responses on the adjacent lines. Although lumped element approximations of high-speed VLSI interconnects are presently adequate for the design processes, the next generation of computers will require an accurate, field theory analysis that includes the effects of coupling, dispersion, and losses. Despite the need for a fast and accurate analysis of high-speed, high-density digital interconnects, there has been very little published on the transient analysis of multi-layer, multi-conductor, high-speed interconnects.

The analysis of VLSI interconnects begin with the characterization of coupled microstrip lines. Pulse distortion on symmetric, coupled, lossy microstrips [1] and the transient parameters of a symmetric five-microstrip line structure [2] have been studied using the spectral domain approach (SDA) in the frequency domain and an inverse-Fourier transform to obtain time-domain results. Distortion of signals on multiple strips has been examined using an impedance matrix approach with line parameters determined by approximate formulas [3]. It has also been shown that it is possible to reduce crosstalk between adjacent lines by placing a third line between the two that is grounded at both ends [4], [5]. Although this method can reduce crosstalk, its effectiveness is dependent on the thickness of the substrate, with greater reduction of crosstalk possible with thicker substrates [4]. In addition, the extra grounded line increases the fabrication costs, makes the inter-line spacing greater than two center conductor widths, and requires via holes to short the ends of the additional line, thus adding stray inductance to the circuit. Also, while direct crosstalk is reduced, the return voltage at the generator end of the driven line is significantly increased, which can couple to the non-driven line and increase the overall amount of crosstalk [4].

An alternative approach, discovered by the authors, retains the small inter-line spacing and controls coupling and crosstalk through the choice of the electrical parameters of the substrates in a multi-layer structure [6]. In the optimum case, it is possible to achieve close to 100 percent reduction in crosstalk and to essentially eliminate distortion due to coupling. The key to this design procedure is to use more than one substrate layer, where the dielectric constant of the upper-most substrate is much greater than the dielectric constant of the lower-most substrate. The proper choice of substrate heights and dielectric constants makes the phase velocities of the even and odd modes essentially equal over an extremely wide bandwidth. This method also works well for very small line spacings, even those less than one center conductor width. Although this new low-coupling structure shows great promise, there is still little information available in the open literature concerning the design methodology and performance of these types of structures.

This paper addresses the problem of high-speed signal propagation on tightly coupled transmission lines and the reduction of crosstalk and coupling distortion through substrate compensation. The characteristics of substrate-

Manuscript received March 22, 1991; revised June 28, 1991.
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IEEE Log Number 9102780

0018-9480/91$01.00

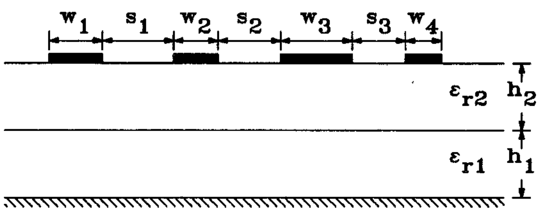

Fig. 1. Geometry of an asymmetric multi-layer, multi-conductor interconnect.

compensated, low-coupling structures are presented in this paper for general multi-layer, multi-conductor structures. The ϵ_{reff} of multi-layer, asymmetric coupled microstrips is given for various substrate height ratios. Pulse propagation on multi-conductor transmission lines is studied, and reduction of crosstalk and coupling distortion on these lines through the use of substrate compensation is proposed. The maximum allowable value of the dielectric constant of the lower substrate is determined for a wide variety of structures. Pulse distortion on a substrate-compensated structure is presented for different inter-line spacings with the same substrate design. This shows that it is possible to achieve significant reduction of both crosstalk and coupling for a wide variety of conductor configurations using the same two-layer substrate geometry.

Full Wave Analysis

The use of the SDA has been well documented in the open literature, and so the technical details are omitted here. A very good explanation of the method can be found in [7] and a simple recursive formulation for the spectral-domain input impedances of structures with multiple substrates and/or superstrates is given in [6]. The geometry for asymmetric multi-layer, multi-conductor interconnects is shown in Fig. 1. Two substrates are shown in the figure, although any finite number of substrates can be easily considered using the recurrence formulation from [6]. In this investigation, the substrates are assumed to be lossless and isotropic, and the conductors have zero thickness and are perfectly conducting. The effects of dielectric loss can be included in the spectral-domain formulation [8], but these losses do not affect the effective dielectric constant when the loss tangent is much less than unity.

The expansion functions for the current densities used in this paper are similar to those proposed by Kobayashi and Iijima [9] and are given by

$$J_{zn}(x) = \frac{T_n(2x/w)}{\sqrt{1-(2x/w)^2}} \tag{1}$$

$$J_{xn}(x) = U_n(2x/w)\sqrt{1-(2x/w)^2} \tag{2}$$

for $n = 0, 1, 2, \cdots$ and $|x| \leq w/2$. $T_n(x)$ and $U_n(x)$ are the Chebyshev polynomials of the first and second kind, respectively, as given in [10]. Although the current distributions of isolated microstrips for the longitudinal (transverse) currents are even (odd) symmetric, for multiconductor lines the current distributions may have both even and odd components. Thus both even and odd symmetric Chebyshev polynomials are used for the current expansion functions. The Fourier transforms of these expansion functions are given in closed form by

$$\tilde{J}_{zn} = (-j)^n \frac{\pi w}{2} J_n\left(\frac{\beta_x w}{2}\right) \tag{3}$$

$$\tilde{J}_{xn} = (-j)^n \frac{\pi(n+1)}{\beta_x} J_{n+1}\left(\frac{\beta_x w}{2}\right) \tag{4}$$

where $J_n(\beta_x)$ is the Bessel function of the first kind of order n. Results obtained with this method have been compared and agree very closely with results presented in the literature [11], [12].

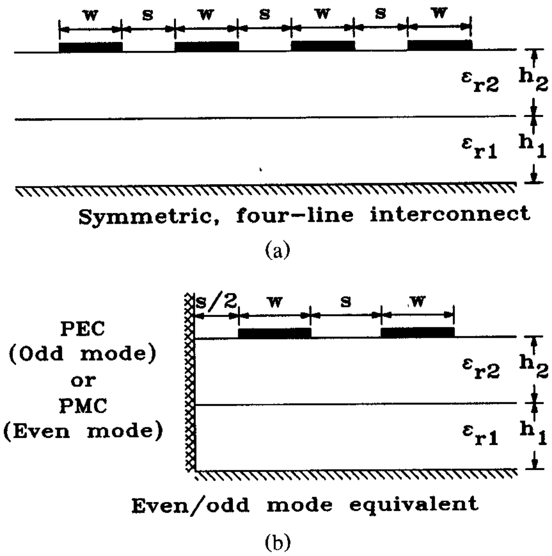

Fig. 2. Symmetric, four-line interconnect and even/odd mode equivalent structure.

When symmetric coupled microstrips are considered, the even/odd mode approach is quick and simple to use. However for asymmetric geometries or those where there are more than two conductors, a different analysis must be used. As an example, consider the symmetric four-line structure shown in Fig. 2(a). The structure can be initially split into even- and odd-symmetric structures by placing a perfect magnetic conductor and a perfect electric conductor, respectively, at the plane of symmetry. However, once this is done, the resulting structure, shown in Fig. 2(b), no longer has any symmetry properties which can be used, and it must be solved in a manner similar to that of asymmetric coupled lines. In the spectral domain, this is accomplished by using independent sets of expansion functions for each conductor, and computing the coefficients of the current densities after ϵ_{reff} has been determined.

Each additional conductor introduces a new propagating mode; i.e., one that has a zero cutoff frequency. Each of these modes has a propagation and an attenuation constant associated with it, although they may be the same as those of other modes. Thus the modes are not uniquely determined by their propagation constants. However, each mode does have a distinct set of current distributions that can be used to uniquely identify the mode. For example, in the low-coupling structures described in [6], the even and odd modes in a two-layer, symmetric coupled microstrip are made to have the same propagation constants by choosing an appropriate combination of substrate heights. However, the two modes can be easily distinguished because the currents on the conductors for the even mode have equal amplitude and are in phase, whereas for the odd mode the currents have equal amplitude, but are 180 degrees out of phase.

For the asymmetric two-conductor case, the current densities for the c and π modes are in-phase and out-of-phase, respectively, just as with the even and odd modes. However, the asymmetric case differs in that the amplitudes of the current densities on the conductors are not the same. In this case, not only the signs, but also the relative amplitudes of the currents are necessary to characterize the mode. Thus to completely characterize an N conductor system, it is necessary to determine N different propagation constants and an N by N matrix that specifies the relative magnitudes of the current densities on each of the N conductors for each of the N modes. In addition, both the matrix and the propagation constants may vary as a function of frequency, and so it is necessary to compute them over a wide spectrum in order to accurately compute pulse distortion.

For a given mode number, say j, the currents on lines $i = 2, 3, \cdots, N$ can be related to the current on line 1 by

$$I_{1j} = a_{ij}(\omega) I_{ij} \tag{5}$$

where I_{ij} is the current on line i for mode j and $a_{ij}(\omega)$ is a frequency dependent proportionality constant. Using (5) it is possible to define a matrix, $[A](\omega)$, that relates the current densities on all N lines for each of the N modes at a given frequency. This N by N matrix is given by

$$[A](\omega) = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ a_{21}(\omega) & a_{22}(\omega) & \cdots & a_{2N}(\omega) \\ \vdots & \vdots & & \vdots \\ a_{N1}(\omega) & a_{N2}(\omega) & \cdots & a_{NN}(\omega) \end{bmatrix}. \tag{6}$$

In order to analyze pulse distortion on N asymmetric coupled lines, we want to find a linear combination of the N modes that gives a unit amplitude signal on one line and no signal on the other lines. This linear combination of the modes shows how the presence of the adjacent lines distorts the input pulse as well as how crosstalk is created on the adjacent lines. For each of the N lines that can be excited, there is a unique linear combination of the N modes that will reproduce the signal on that line with no signal on the other lines. The coefficients of the linear combination of the N modes that satisfies the above requirement can be found by solving the following matrix equation:

$$[A](\omega) \mathbf{c}_k(\omega) = \mathbf{e}_k \tag{7}$$

which is given in an expanded form as

$$\begin{bmatrix} 1 & 1 & \cdots & 1 \\ a_{21}(\omega) & a_{22}(\omega) & \cdots & a_{2N}(\omega) \\ \vdots & \vdots & & \vdots \\ a_{N1}(\omega) & a_{N2}(\omega) & \cdots & a_{NN}(\omega) \end{bmatrix} \begin{bmatrix} c_{1k}(\omega) \\ c_{2k}(\omega) \\ \vdots \\ c_{Nk}(\omega) \end{bmatrix} = \begin{bmatrix} e_{1k} \\ e_{2k} \\ \vdots \\ e_{Nk} \end{bmatrix} \tag{8}$$

where $c_{jk}(\omega)$ is the unknown constant of the linear combination for the jth mode with the kth conductor being excited and $\mathbf{e}_k$ is the standard unit vector which is defined as

$$e_{ik} = \begin{cases} 0 & \text{for} \quad i \neq k \\ 1 & \text{for} \quad i = k. \end{cases} \tag{9}$$

Since the $a_{ij}(\omega)$'s will normally change with frequency, the c_{jk}'s will also vary and therefore (8) must be solved repeatedly to find the appropriate linear combinations for all the necessary frequencies in the pulse spectrum.

Because there are N possible conductors which may be excited, and therefore N distinct linear combinations, it is possible to write the solution for all possible linear combinations as an N by N matrix, $[C](\omega)$. This matrix is obtained by gathering all of the column vectors $\mathbf{c}_k(\omega)$ in (8) for $k = 1 \rightarrow N$ into a single square matrix, which results in

$$[A](\omega)[C](\omega) = [I] \tag{10}$$

where $[I]$ is the identity matrix. Thus the set of linear combinations for all possible unit excitations of a single line, $[C](\omega)$, is simply the inverse of the current coefficient matrix, $[A](\omega)$. So not only are all of the linear combinations distinct, they are also mutually orthogonal.

This approach simplifies to that of the even/odd mode and c/π mode approach when the number of conductors is only two. For the even/odd mode approach, (10) can be written more simply as

$$\begin{bmatrix} c_{e1} & c_{e2} \\ c_{o1} & c_{o2} \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ a_{2e} & a_{2o} \end{bmatrix}^{-1} \tag{11}$$

where the mode numbers have been replaced with e for the even mode and o for the odd mode. Since the structures for the even/odd mode approach are symmetric, a_{2e} and a_{2o} are known for all frequencies to be 1 and -1, respectively. Solving the above equation, we find that $c_{e1} = c_{o1} = c_{e2} = \frac{1}{2}$ and $c_{o2} = -\frac{1}{2}$. In matrix form, the linear combinations are given as

$$[C](\omega) = [A]^{-1}(\omega) = \begin{bmatrix} \frac{1}{2} & \frac{1}{2} \\ \frac{1}{2} & -\frac{1}{2} \end{bmatrix}. \tag{12}$$

Thus if one-half of the even mode excitation is added to one-half of the mode mode excitation, this will give a signal on line 1, with none on line 2. Likewise, if one-half of the odd mode excitation is subtracted from one-half of the even mode excitation, then the signal will be on line 2 with none on line 1. These results are the same as those obtained using the even/odd mode approach. Note that for the even/odd mode approach, the current coefficient matrix $[A](\omega)$, and hence $[C](\omega)$, is constant for all frequencies.

Equation (8) can also be applied to asymmetric, two conductor structures by using c and π modes. In this case, a_{2c} and $a_{2\pi}$ are not known beforehand, since they depend on the geometry, the electrical parameters of the system, and the frequency. Replacing the mode numbers 1 and 2 with c and π, respectively, the matrix equation is now given by

$$[C](\omega)=\begin{bmatrix} c_{c1} & c_{c2} \\ c_{\pi 1} & c_{\pi 2} \end{bmatrix}=\begin{bmatrix} 1 & 1 \\ a_{2c} & a_{2\pi} \end{bmatrix}^{-1}. \quad (13)$$

This has a simple, closed form solution which results in

$$[C](\omega)=\begin{bmatrix} \dfrac{a_{2\pi}}{a_{2\pi}-a_{2c}} & \dfrac{-1}{a_{2\pi}-a_{2c}} \\ \dfrac{-a_{2c}}{a_{2\pi}-a_{2c}} & \dfrac{1}{a_{2\pi}-a_{2c}} \end{bmatrix}. \quad (14)$$

Once $[C](\omega)$ has been determined for all the relevant frequencies, the next step is to compute the time domain response for each of the N modes using the inverse Fourier transform. The pulse spectrum in each integrand is multiplied by both the current amplitude coefficient, $a_{ij}(\omega)$, and by the coefficient of the appropriate linear combination, $c_{jk}(\omega)$. The total response for each of the N lines is then obtained by summing the responses of all N modes. Thus, the response on line i to a signal input on line k at a time t and distance z may be calculated from

$$v_i(t,z)=\frac{1}{2\pi}\sum_{j=1}^{N}\int_{-\infty}^{\infty} a_{ij}(\omega)c_{jk}(\omega)\tilde{V}(\omega)e^{j[\omega t-\gamma_{zj}(\omega)z]}\,d\omega \quad (15)$$

where $\gamma_{zj}(\omega)=\alpha_{zj}(\omega)+j\beta_{zj}(\omega)$ is the complex propagation constant of the jth mode.

Results

The goal in the design of a substrate-compensated, low-coupling structure is to equalize the phase velocities of the independent modes by adjusting the relative dielectric constants of the substrates and/or the heights of the substrates. Throughout this investigation, the total substrate height, $h_{\text{tot}}=h_1+h_2$, is held constant and we consider as parameters the frequency, spacing, $\epsilon_{r1}, \epsilon_{r2}$, and the height ratio, h_1/h_{tot}. Also, only open structures are presented here, but the same procedure may be used to consider structures with a cover sheet and/or side walls.

While the characteristics of both asymmetric and symmetric coupled microstrips are well known for single substrates, the properties can change substantially when more than one substrate is used. For example, consider

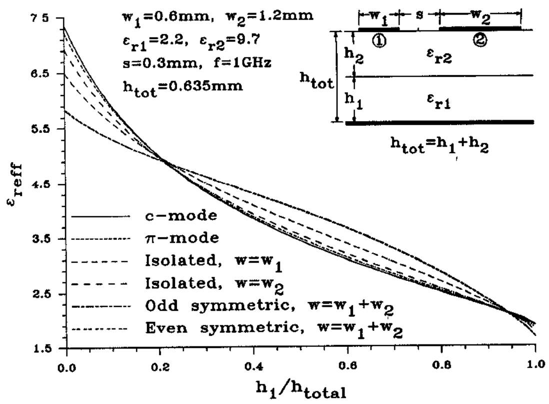

Fig. 3. The ϵ_{reff} of asymmetric coupled microstrips and single isolated microstrips on a two-layer substrate versus the height ratio, h_1/h_{tot}.

the two-substrate, asymmetric coupled line structure shown in Fig. 3. The effective dielectric constant of the c (in phase) and π (out of phase) modes, as a function of the height ratio, h_1/h_{tot}, are displayed in Fig. 3. Also included is the ϵ_{reff} of a single isolated strip of width w_1 and one of width w_2, which are the limiting cases for the c and π modes as the spacing is increased. As the inter-line spacing approaches to zero, the limiting case for the c mode is a strip of width w_1+w_2 with an even symmetric current distribution, and for the π mode it is a strip of width w_1+w_2 with an odd symmetric current distribution. In Fig. 3, the ϵ_{reff} for the odd-symmetric isolated strip, $w=w_1+w_2$, is essentially the same as the ϵ_{reff} of the π mode, as would be expected for a structure with very small inter-line spacing. Likewise, the ϵ_{reff} for the even-symmetric isolated strip, $w=w_1+w_2$, is very close to that of the c mode.

For single substrate structures, ϵ_{reff} increases as the width of the microstrip is increased. However, as shown in Fig. 3, there are values of the height ratio for which the thinner strip, $w=w_1$, has larger ϵ_{reff} than the wider strip, $w=w_2$. Also note, that while the c mode has a higher ϵ_{reff} than the π mode for single substrate structures, for some values of the height ratio, this is reversed and at two points the ϵ_{reff}'s of the two modes are equal. This is similar to the case of symmetric coupled microstrips on substrate-compensated low-coupling structures [6] where the even and odd mode phase velocities are equalized through the proper choice of the electrical parameters of the substrate. Thus it is possible to design low-coupling structures using asymmetric-coupled lines as well as by using symmetric-coupled lines.

An example of a multi-conductor, symmetric interconnect is shown in Fig. 4 where the ϵ_{reff} of the four independent modes is plotted as a function of the height ratio. The relative signs of the currents on the four conductors is shown for each of the four modes in the graph titles in Fig. 4. The sign of the current on the first conductor is arbitrarily chosen to be positive, and the signs of the currents on the other conductors are determined from the spectral-domain solution. As with the

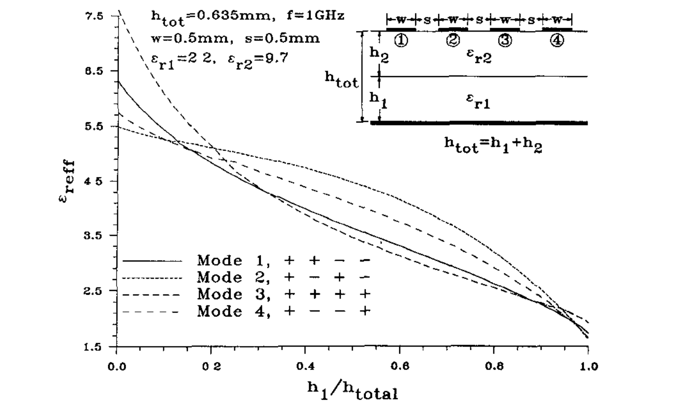

Fig. 4. The ϵ_{reff} for the four modes of a symmetric coupled four-line structure on a two-layer substrate versus the height ratio h_1/h_{tot}.

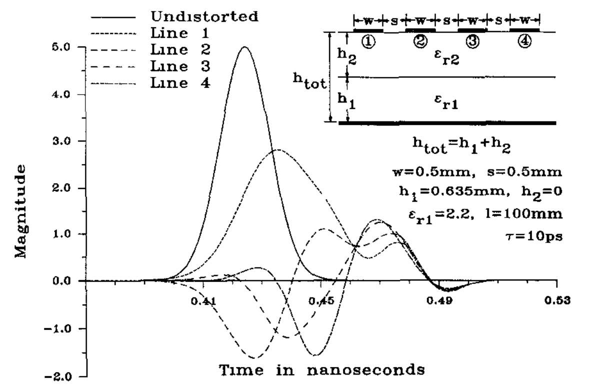

Fig. 5. Pulse distortion on four-line symmetric coupled interconnects with a single substrate layer.

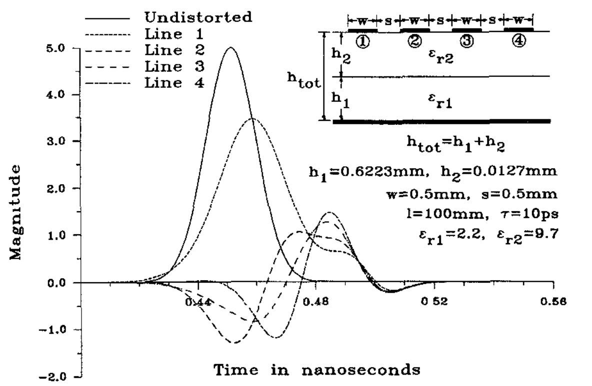

Fig. 6. Pulse distortion on four-line symmetric coupled interconnects using substrate compensation.

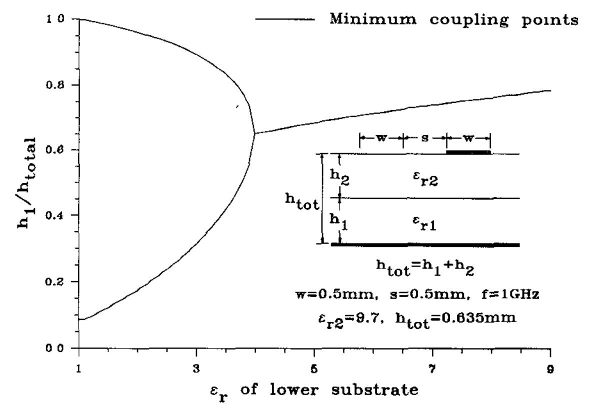

Fig. 7. Minimum coupling points versus ϵ_{r1} for a substrate-compensated low-coupling structure.

symmetric and asymmetric two-conductor cases, the ϵ_{reff} of each of the modes in the four-conductor case changes at different rates as the height ratio is varied. Thus while mode 3 has the highest ϵ_{reff} on single layer structures, for height ratios between 0.3 and 0.85 it has the lowest ϵ_{reff}. However, unlike the two-conductor cases, it may not be possible to find a height ratio where all four modes have exactly the same ϵ_{reff}. On the other hand, it is possible to choose a height ratio where the differences between the ϵ_{reff}'s of the four modes is minimized.

Pulse propagation on a symmetric four-conductor structure with a single substrate, $\epsilon_r = 2.2$, is shown in Fig. 5 with the dimensions of the structure identified in the figure. A Gaussian pulse with a half-width, half-maximum of 10 ps is sent down line 1 and the time-domain response on all four lines is plotted at a distance of 100 mm. At this distance, the amplitude of the intended pulse has been reduced by 40 percent, and the pulse has been significantly widened. In addition, the amplitude of the responses on the adjacent lines has risen to 30 percent of the amplitude of the original signal, which represents a significant amount of crosstalk.

To improve the intended signal and reduce the crosstalk on the adjacent lines, a thin upper substrate is introduced, with a relative dielectric constant that is much higher than that of the lower substrate. An upper substrate with $\epsilon_{r2} = 9.7$ and thickness of $h_2 = 0.0127$ mm is placed on a lower substrate of $\epsilon_{r1} = 2.2$ and thickness $h_1 = 0.6223$ mm. The same Gaussian pulse from Fig. 5 is sent on line 1, and the resulting signals on all four lines for the same distance, $l = 100$ mm, is shown in Fig. 6. Using this configuration, the intended signal suffers only a 30 percent reduction in amplitude, and the widening of the pulse is reduced as well. The crosstalk on the adjacent lines is also reduced using this structure.

One important characteristic of a low-coupling structure is the value of ϵ_{r1} which equalizes the even and odd mode phase velocities. In Fig. 7 the value of h_1/h_{tot} which either minimizes the difference between the modal phase velocities or makes them exactly equal is plotted as a function of ϵ_{r1} with $\epsilon_{r2} = 9.7$ for the structure shown. For $9.7 \geq \epsilon_{r1} \geq 4.0$, there is one value of the height ratio that minimizes the difference between the modal ϵ_{reff}, but for which they are not equal. However, at $\epsilon_{r1} \approx 4.0$ the graph splits because there are now two values of the height ratio for which the modal phase velocities are exactly equal. The value of ϵ_{r1} at which this split occurs is the maximum allowed value of ϵ_{r1} for which the phase velocities can be exactly equalized.

In Fig. 8 the maximum value of ϵ_{r1} which allows the equalization of the modal phase velocities is plotted as a function of ϵ_{r2}. Two different strip widths, each with

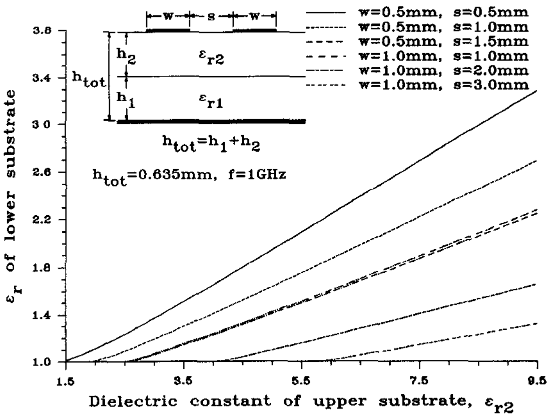

Fig. 8. Maximum ϵ_{r1} which can be used to design a zero-coupling structure versus ϵ_{r2} for various conductor geometries.

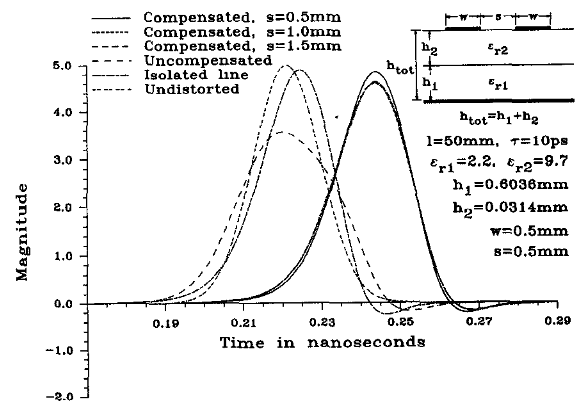

Fig. 9. Pulse distortion on symmetric coupled microstrips for various line spacings, signal line response.

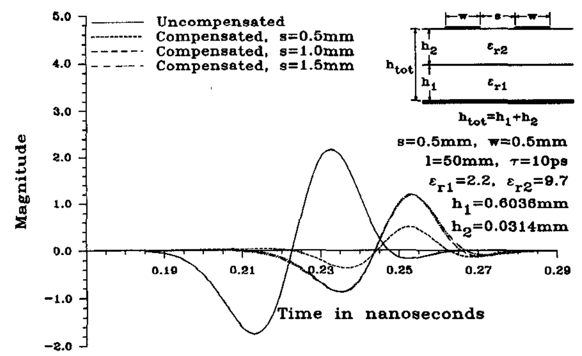

Fig. 10. Pulse distortion on symmetric coupled microstrips for various line spacings, sense line response.

three different spacings, are considered. All values of ϵ_{r1} below a given curve can be used with that conductor configuration to create a zero-coupling structure; i.e., one where the modal phase velocities are exactly equal. For example, if $w = 0.5$ mm, $s = 0.5$ mm, and $\epsilon_{r2} = 5.5$, then values of $\epsilon_{r1} \leqslant 2.1$ allow the design of structures with zero coupling. However, note that the coupling can still be greatly reduced, but not eliminated, by using for the lower substrate an ϵ_{r1} which is somewhat greater than the maximum allowable value. The height ratio is then chosen to be that which minimizes the difference in the modal phase velocities, as in Fig. 7.

As the spacing between the center conductors is increased, as shown in Fig. 8, the maximum allowable value of ϵ_{r1} decreases. However, as the spacing increases, the overall coupling between the two lines decreases and the restriction on ϵ_{r1} becomes less critical. When the width of the center conductor is increased, the maximum allowable value of ϵ_{r1} decreases. All of the graphs show a somewhat linear dependence. Therefore it may be possible to determine simple, curve-fitted formulas that can be used in a design process.

Another important parameter of a low-coupling structure is the inter-line spacing. For a given spacing and width on a low-coupling structure, there are two values of the height ratio which eliminate coupling and crosstalk at a given frequency. As the spacing changes, the values of this height ratio change as well. Thus it is interesting to consider the following question: if a structure were designed to have zero-coupling for a certain center conductor width and spacing, what would the effect be on the coupling and crosstalk for lines with different spacings on the same substrate combination? To answer this question, a two-substrate, symmetric coupled microstrip is considered with a total substrate height of $h_{\text{tot}} = 0.635$ mm, $\epsilon_{r1} = 2.2$, $\epsilon_{r2} = 9.7$, a center conductor width of 0.5 mm, and spacing of 0.5 mm. The height of the lower substrate is chosen to be $h_1 = 0.6036$ mm, which equalizes the even and odd mode phase velocities in the quasi-static frequency range. Pulse distortion on this structure is then considered for line spacings of $s = 0.5$, 1.0, and 1.5 mm. These results are compared to the undistorted pulse, a pulse on an isolated line of the same width, and a pulse on symmetric coupled lines of the same widths, with a spacing of 0.5 mm. All three of the latter cases are computed for a single substrate of $\epsilon_r = 2.2$ and height 0.635 mm.

The time domain responses for all six cases are shown in Fig. 9 for the signal or intended line and in Fig. 10 for the sense or adjacent line. A Gaussian pulse with a half-width, half-maximum of 10 ps is used and the transient response for both lines is taken at a distance of 50 mm. The amplitude of the signal line response for the uncompensated line has been significantly reduced from the amplitude of the undistorted pulse and it has almost doubled in width. On the other hand, the pulse on the compensated substrate with a spacing of 0.5 mm has almost no degradation of amplitude nor has it widened noticeably. The signal line responses of the pulses on compensated substrates with spacings of 1.0 and 1.5 mm show a little loss of amplitude due to coupling distortion, but it is very slight.

The sense line response for the uncompensated and compensated structures is shown in Fig. 10. The amplitude of the sense line response for the uncompensated structure is almost 50 percent of the amplitude of the undistorted pulse, representing a significant amount of crosstalk. The amplitude of the sense line response for the compensated substrate with a spacing of 0.5 mm is only 15 percent of the amplitude of the original pulse, much less than that of the uncompensated line. For the other two spacings, the amplitude of the sense line response is about 22 percent of the amplitude of the undistorted pulse, which is still less than one-half of that of the uncompensated line.

Note that as the spacing increases beyond 1.5 mm, the coupling and crosstalk will be reduced, since the coupling varies inversely with the spacing. Thus it is possible to choose two dielectrics and a height ratio for which the coupling and crosstalk are significantly reduced for a wide range of center conductor spacings. This allows a great deal of flexibility in the design of the low-coupling structures for high-speed interconnects. Many different conductor geometries can be used on the same substrate combination with the crosstalk being significantly reduced for all of the geometries.

Conclusion

The faster speeds and small inter-line spacings being considered for future interconnect designs can result in unacceptable levels of crosstalk and significant distortion of the intended signal due to coupling. Some of the characteristics of substrate-compensated, low-coupling structures are investigated for symmetric and asymmetric coupled lines as well as for a symmetric structure with four lines. Pulse propagation on the symmetric four-line structure was presented, showing how coupling and crosstalk can be reduced through proper choice of the substrate parameters. Pulse distortion on substrate-compensated, symmetric coupled microstrips is also considered for various inter-line spacings, showing that significant reduction of crosstalk and coupling can be achieved for a wide variety of conductor geometries using the same substrate combination. By designing high-speed interconnects on substrate-compensated, low-coupling structures, it will be possible to achieve an extremely high density of signal conductors, using inter-line spacings of less than one center conductor width, while keeping crosstalk and coupling distortion to a minimum

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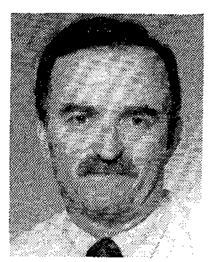

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