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Dielectric Road Waveguide by Using Wave Band MM and Sub Band MM in Antenna Technology

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Abstract:

Design of an all-electrical and easy to package wideband chip-to-chip solution on a multi-mode dielectric waveguide is discussed. Di®erent parameters such as bandwidth-per-pitch, range, as well as crosstalk in aggregated lines are analyzed. Using a Rogers RO3006 material, a bandwidth-per-pitch of 16 GHz/mm (and an absolute bandwidth-per-line of 2 £ 40 GHz) at the center frequency of 100 GHz is achieved while maintaining a range of 1 m with crosstalk below 15 dB. The sensitivity of the line to the curvature is also examined. The signal is coupled from the silicon chip to the degenerate fundamental and polarization-orthogonal E_x^{11} and E_y^{11} waveguide modes using planar electric and slot dipole antennas, respectively. The large available bandwidth will be channelized in frequency for optimal overall e±ciency and throughput with a CMOS transceiver. The performance sensitivity of the structure to possible fabrication imperfections is examined and discussed. The proposed waveguide o®ers a solution for Tera bit-per-second (Tbps) fully-electrical wireline links.

1. INTRODUCTION:

There is a large demand to substantially increase overall system bandwidth as well as connectivity at all levels- from individual ICs to backplanes.

As aggregate link data rate requirements are approaching Tbps levels, current physical channels are nearing their limits. To this end, a power and data-rate \wall" is emerging on the horizon. This is in terms of maximum \data °ux density" (bandwidth per unit width) and link e±ciency (in terms of energy per bit) when the bandwidth of conventional planar on-board transmission lines is pushed to tens of GHz. In assessing communication channels, several ¯gures of merit exist: bandwidth and data-rate (considering width and length of the channel), overall energy e±ciency, and the associated cost (e.g., with extra components or need for precision/complex packaging). In improving data bandwidth, main culprits are the high-loss, dispersion, crosstalk and re°ections in the conventional planar TEM electrical channel. In these lines, the capacity of the link is primarily limited by skin e®ect and impedance mismatches [1]. For medium to large distances and/or to achieve high data rates, the cost of equalization in electrical lines becomes excessive. Optical ¯ber links have been extensively used in long-haul communication and are now also slowly making their way into the short-range arena. However, the associated power and cost overhead (due to precision packaging and the addition of electro-optical modules) pose limitations on adaptability for ubiquitous board-level interconnects.

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In this paper, we propose a structure that addresses these concerns. An all-electrical, easy to package and planar structure on a dielectric waveguide is presented and di®erent aspects (as an interconnect) are analyzed. This has applications for chip-to-chip interconnects as well as for short active cables. Dielectric waveguides made of simple dielectric strips are known for very low transmission loss at mm-wave and sub-THz frequencies [2, 3]. The challenge lies in designing an e±cient, wideband and low-cost coupling structure that makes appropriate use of the waveguide modes. Here, we propose a novel planar feed structure to excite two *polarization- orthogonal* modes of the waveguide, namely the E_x^{11} and E_y^{11} modes. This doubles the applicable bandwidth of the dielectric waveguide without compromising performance. Planar electric dipole and slot dipole antennas can launch the E_x^{11} and E_y^{11} modes, respectively. Excitation with planar feed antennas simpli¯es the assembly and enables the use of conventional packaging with CMOS transceivers. Fig. 1 shows a conceptual schematic of the proposed structure. As depicted in Fig. 1(b), the electric ¯eld orientations of the ¯rst two modes are perpendicular to each other.

2. DIELECTRIC WAVEGUIDE DESIGN:

Dielectric waveguide for data communication has been proposed in [4, 5]. In [5], due to the se-lected structure and modes, the implementation presents lower coupling e±ciency to the waveguide as well as strong waveguide mode-coupling and corresponding length dependencies that a®ect ro-bustness. Mode leakage due to discontinuities may also pose limitations. Our approach focuses on fundamental and polarization-orthogonal degenerate modes where the mode-coupling to higher order modes is greatly suppressed and the coupling e±ciency (from the chip to the waveguide) is considerably better. It should be noted that another communication channel based on substrate-integrated waveguide (SIW) is reported in [6]. Due to operating at lower mm-wave frequencies, the bandwidth of the structure is limited.

As shown in Fig. 1, the proposed waveguide is simply made of a dielectric strip, dimensions of which are primarily determined by the operating frequency band of the ¯rst two fundamental E_x^{11} and E_y^{11} modes. In addition, limiting the cross section avoids mode cross coupling along the waveguide by eliminating higher order modes. The ¯eld con¯nement in the cross section of the waveguide is related to the dielectric strip permittivity as well as waveguide dimensions. Higher permittivity leads to an increased ¯eld concentration within the waveguide core, as opposed to ¯elds on the outside of the strip, allowing for a smaller pitch. In order to generalize the argument, the available bandwidth-per-pitch and the maximum length of waveguides with high and low permittivities are approximated for di®erent frequency bands. It should be noted that dielectric materials applicable for millimeter-waves usually have a permittivity in the range of 2 to 12. In [7] the dispersion diagram of square rectangular dielectric waveguides made of materials with permittivity of 12 and 2.3 are extracted in terms of a normalized frequency parameter ($v = k_o a\sqrt{\epsilon_r - 1}$). The values of v at cut-o® frequencies are known. Therefore the bandwidth can be calculated as a function of center operating frequency:

$$BW = \frac{Kc}{2\pi a^2\sqrt{\epsilon_r - 1}} \quad (1)$$

$$a = \frac{4.5c}{2\pi f_o^2\sqrt{\epsilon_r - 1}} \quad (2)$$

where c, f_o, and a are the free space wave velocity, center frequency, and waveguide cross section, respectively. K is 1.5 and 2, respectively, for dielectrics with permittivity of 12 and 2.3. Assuming a pitch of $2a$ for the waveguide, the bandwidth-per-pitch parameter is plotted in Fig. 2. This normal-ized bandwidth for a single mode is proportional to f_o^2. In addition, with increasing the permittivity the bandwidth-per-pitch increases. This is mainly due to the increasing ¯eld con¯nement. On the other hand, the maximum operating frequency is restricted by the attenuation of the waveguide.

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An approximate relation for the attenuation constant of a dielectric waveguide is presented in [8]. Assuming a maximum acceptable transmission loss of 50 dB for the link, the maximum length of the waveguide is calculated for dielectric loss tangent of 0.001 and plotted in Fig. 2. As the frequency increases, due to higher attenuation, the maximum available length (L_{max}) decreases. In addition, for the same loss tangent, the attenuation per unit length increases with higher permittivity. It should be noted that dielectrics with higher permittivity usually also present a higher loss tangent. From this analysis, the required length of the link determines the center frequency, which in turn limits the maximum bandwidth of the link. Utilizing repeaters can partially alleviate this tradeo®.
In a multiplexed lines scenario, the minimum pitch of single wires is determined by the maximum

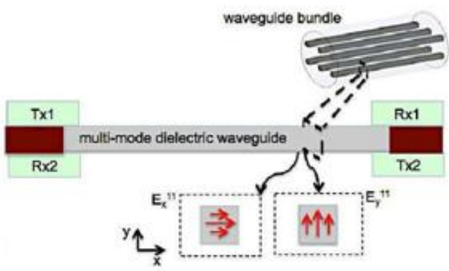

Figure 1: Conceptual schematic of the proposed multi-mode dielectric waveguide for chip-to-chip in-terconnects or active cables.

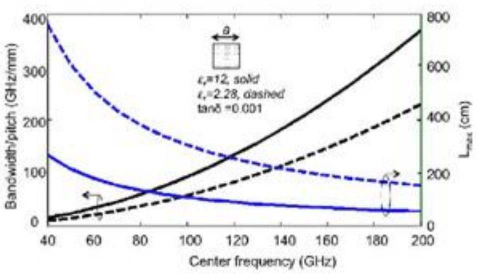

Figure 2: Bandwidth-per-pitch and maximum length of the link (with acceptable transmission loss of 50 dB) in terms of operating frequency.

Acceptable crosstalk, which is related to the length of the link as well as the ¯eld expansion around the waveguide.

Fig. 3 compares the expansion of E_x component of E_x^{11} mode in the cross section of waveguides at the lower edge of the operating frequency band (with lower ¯eld con¯nement) for di®erent permittivities. The discontinuity in the E_x along x-axis is due to discontinuity in permittivity seen by normal electric ¯eld. The E_y^{11} mode ¯eld distributions are dual of the E_x^{11} mode. As observable, ¯elds of waveguide with lower permittivity propagate with a relatively lower con¯nement in the cross section. This limits the minimum distance between possible parallel \wires" in a multiplexed link. In order to quantify this argument, the maximum range of a link with a crosstalk of less than ¡15 dB is calculated and shown in Fig. 4. As expected, for the same length, links composed of waveguides with lower permittivity need a larger pitch. Taking this analysis into account, a rectangular strip waveguide made of Rogers RO3006 dielec-tric material (loss tangent 0.002) with a medium permittivity of 6.15 is constructed and designed for a 1m link.

Assuming an acceptable transmission loss of 50 dB (including waveguide attenuation and mode excitation losses) the appropriate operating frequency is calculated. Simulations show when operating at frequencies around 100 GHz (waveguide cross section: 0:85£0:85 mm^2) an attenuation of 0.044 dB/mm (44 dB/m) is achievable. This waveguide supports a bandwidth of approximately 2 £ 40 GHz each from 80 GHz to 120 GHz. These waveguides can then be bundled together for a larger aggregate throughput. Returning to Fig. 4, to have a minimum crosstalk of 15 dB a pitch of larger than 1:65$¸_0$ ($¸_0$ is the free-space wavelength at the center operating frequency) is required. Therefore, a bandwidth-per-pitch of 2 £ 16 GHz/mm is achievable. The sensitivity of transmission and mode cross coupling to the bending of the\wire" (see Fig. 5) is also investigated. Simulations show an extra loss of less than 0.2 dB at a 90$^{\pm}$ bend with radius of 5 mm at 100 GHz. The amount of mode cross coupling at such a bend is negligible and less than ¡33 dB in the whole frequency range from 80 GHz to 120

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GHz. On the transceiver front, several wideband millimeter-wave silicon IC solutions have been realized in CMOS and BiCMOS technologies. We have recently demonstrated a silicon transceiver with > 30 GHz of bandwidth (on a 94 GHz carrier) withon-chip antennas [9]. This chip has su±cient bandwidth to drive one of the orthogonal channels in the proposed dielectric waveguide. A simple QPSK modulation achieves over 120 Gbps throughput on a single waveguide. Frequency chan-nelization and interleaving can enable this bandwidth in baseband. Other high-speed mm-wave CMOS transceivers for wireless communication have been demonstrated [10, 11]. Here, the channel response is signi¯cantly better than the wireless case.

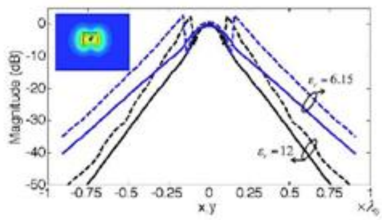

Figure 3: E_x magnitude of the E_x^{11} mode in the waveguide cross section along x-axis (dashed) and y-axis (solid), inset shows the E-¯eld pattern in the cross section.

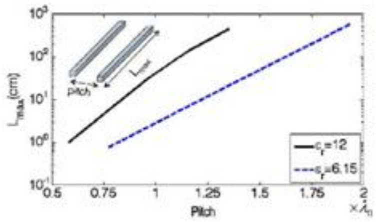

Figure 4: Maximum length of aggregated lines for a crosstalk of less than ¡15 dB for waveguides with permittivity of 12 (solid line) and 6.15 (dashed line).

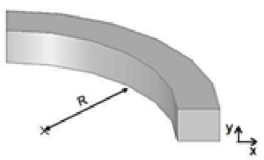

Figure 5: Multi-mode propagation along a bend wit a radius of R = 5 mm.

3. SENSITIVITY ANALYSIS OF THE E_X^{11} AND E_Y^{11} MODELAUNCHERS:

As seen in Fig. 1(b), the E_x^{11} mode is characterized by an electric ¯eld primarily in the x direction and thus, can be excited by an electric dipole along the axis. Conversely, the E_y^{11} mode is the dual of the E_x^{11} mode and as lot dipole antenna (dual of electric dipole) can be utilized to launch the mode into the waveguide. Recently we have presented planar structures, which e±ciently launch the signal from integrated circuits into dielectric waveguides [12, 13]. A drawing of the structure presented in [13] is depicted in Fig. 6. The electric and slot dipoles are printed on 100 ¹m thick RO3010 substrates of permittivity 10.2, and placed on both sides of the waveguide. The high permittivity of the substrate reduces the radiation and therefore enhances the coupling e±ciency. Fundamental design parameters of the excitation structures are detailed in [13]. Transmission parameters of S_{13}, S_{24} and isolation parameters of S_{14} , S_{23} , and S_{12} of the back-to-back structure (see port numbering in Fig. 6) with a length of 10 mm are plotted in Fig. 7.

The obtained results show a 3-dB transmission bandwidth from 88 GHz to 115 GHz for the E_x^{11} mode and from 88 GHz to 120 GHz for the E_y^{11} mode. The coupling e±ciency of each excitation structure is better than 3 dB in the operating frequency band. The bandwidth and e±ciency of the ¯eld coupling can be improved by adding parasitic elements or loading to the dipole or slot antennas (e.g., Yagi-like planar electric and slot dipole antennas). An isolation of better than ¡50 dB between the two input couplers (S_{12}) is also demonstrated. The amount of cross coupling between input electric dipole and output slot dipole (S_{14}) as well as between input slot dipole and output electric dipole (S_{23}) is also better than 50 dB. The sensitivity of coupling e±ciency to the horizontal and angular misalignment of excitation sections is investigated. Fig. 8 shows the e®ect of horizontal o®set of the E_x^{11} excitation section from the center of the waveguide. The overall performance is not highly perturbed and discrepancies are less than 0.5 dB.

The amount of coupling at 100 GHz for di®erent o®sets is plotted in Fig. 9. As the o®set increases the coupling e±ciency decreases. This is mainly due to radiation and partially due to input impedance mismatch. Beside the horizontal o®set, the e®ect of angular misalignments is also examined. Figs. 10 and 11 demonstrate the e®ect of this misalignment on E_x^{11} and E_y^{11} mode launchers, respectively. These results show that E_y^{11} mode is relatively more sensitive to angular misalignment. This may be due to partial excitation of surface wave TM_0 mode in the substrate, which results in radiation in μ direction.

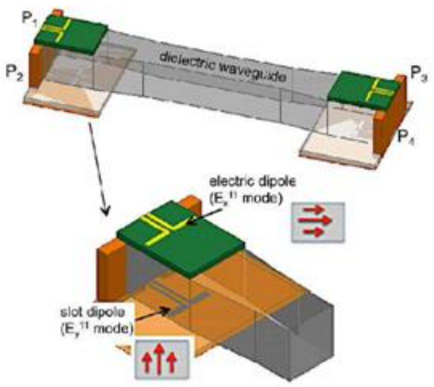

Figure 6: Drawing of the proposed structure for excitation of E_x^{11} (using an electric dipole) and E_y^{11} (using a slot dipole) modes.

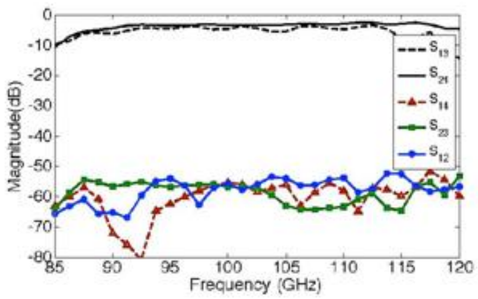

Figure 7: Transmission and isolation parameters of a 10 mm line terminated with E_x^{11} and E_y^{11} mode launchers (see port numbering in Fig. 6).

4. COMPARISON TO OTHER APPROACHES:

Table 1 compares the planar mm-wave dielectric waveguide approach to two other common inter-connects being used for high-data rate chip/board level communication. In [1] the partition length

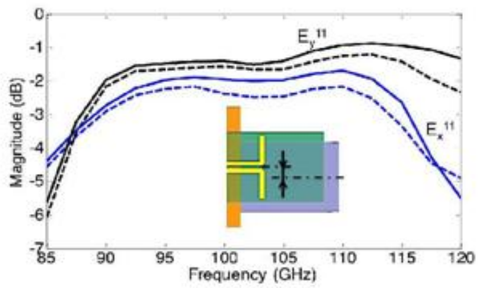

Figure 8: The sensitivity of the coupling e±ciency to the misalignment of the waveguide (solid line: original structure, dashed line: 150 ¹m misalignment).

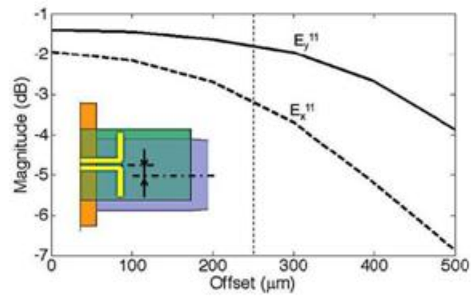

Figure 9: The e®ect of horizontal misalignment of both excitation sections. E_x^{11} is more sensitive to this o®set.

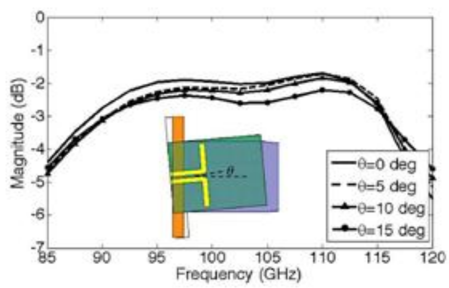

Figure 10: Sensitivity of the coupling e±ciency of the E_x^{11} mode launcher to angular mismatches.

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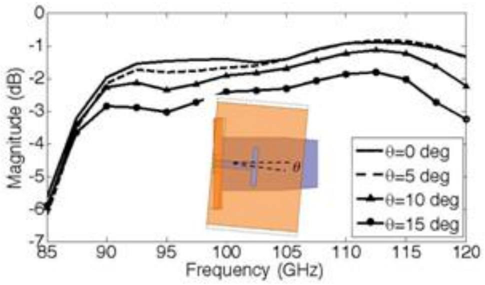

Figure 11: Sensitivity of the coupling e±ciency of the E_y^{11} mode launcher to the angular mismatch.

Table 1: Interconnects options for high-data rate links.

Physical Medium:	Common Electrical	Optical	mm-Wave Waveguide
Transmission loss	High	Very low	Low
BW/pitch	Small	Very large	Large
Crosstalk	High	Very low	Low-Medium
Energy/Bit for high data-rates	High	High	Potentially Low
Overall Cost	Low	High	Medium

Over which board level optical interconnects outperform conductive electrical wire lines is calcu-lated. Common electrical interconnects such as planar microstrip or two-wire lines su®er from high loss and crosstalk. This limits the maximum applicable length and maximum bandwidth-per-pitch of a link in a single or aggregated lines scenarios. In addition, to achieve high throughput in these electrical links, higher order modulation schemes and complicated equalizers are inevitable. This reduces the power e±ciency substantially and increases the cost and complexity of the system. Op-tical interconnects o®er very wide bandwidth, small crosstalk and extremely low transmission loss. Hence they can be used for short/long-range communication. However the cost and complexity due to high precision needed for assembly and electro-optical modules limits the application of such interconnects. Current approach based on a low-loss dielectric waveguide and its multi-mode excitation presents a large bandwidth over a reasonable length (around 1 m). In addition, the all-electrical and planar construction o®ers a low-cost implementation option. Superior channel response leads to simpler equalization circuits and can enhance the power e±ciency of these high-throughput links.

5. CONCLUSION:

Di®erent aspects of an all-electrical, low-loss, and extremely wideband interconnect technology based on dielectric waveguides are analyzed. For a given range and receiver sensitivity various link parameters | the operating frequency, minimum line pitch, and consequently the maximum bandwidth-per-pitch | are calculated and veri¯ed with simulations. A reliable performance of the interconnect under physical curvature of the line is con¯rmed. Millimeter-wave signals are simulta-neously launched into a dielectric waveguide using optimized planar dipole and slot antennas. The sensitivity of coupling e±ciencies to horizontal and angular misalignment of the excitation struc-tures is investigated. This low-cost interface presents a high-speed and low-dispersion alternative to current electrical and optical links. When combined with appropriate frequency channelization and interleaving, these links can achieve aggregate data rates approaching several Tbps over distances as long as one meter.

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