Strong LOS MIMO for Short Range MmWave Communication — Towards 1 Tbps Wireless Data Bus

Xiaohang Song*, Lukas Landau*, Johannes Israel†, and Gerhard Fettweis*

* Vodafone Chair Mobile Communications Systems, Technische Universität Dresden, Dresden, Germany
† Institute of Numerical Mathematics, Technische Universität Dresden, Dresden, Germany
Email: {xiaohang.song, lukas.landau, johannes.israel, fettweis}@tu-dresden.de

Abstract—In this paper, we propose a wireless data bus system design which relies on a strong Line-of-Sight MIMO approach. An analog MIMO equalizer, which equalizes the deterministic MIMO channel is involved. Instead of interference suppression, the analog MIMO equalizer aligns the phases of the received signals and enhances the desired signal while it suppresses the undesired ones simultaneously. Although the magnitudes of different signal components of the received signals are preferred to be unique, further studies are applied in our work to investigate the validation of our system design with non-unique magnitudes by using practical on-board antennas. It is shown that the proposed system works well with non-unique magnitudes where undesired remaining interference occurs with limited power in comparison with the desired ones. Furthermore, the proposed design shows a great potential for putting a 1 Tbps wireless data bus into practical systems with moderate transmit power and fairly simple modulation schemes which provide high energy efficiency.

I. INTRODUCTION

The previous works [1] and [2] showed that beamforming as well as beam switching can be utilized for setting up communication links between computing nodes that are facing each other. However, in order to provide high enough spatial resolution such that independent wireless channels can be set up for every pair of the boards, very large arrays have to be utilized for generating pencil beams. In our approach, independent parallel SISO channels can be generated via a switching network and an analog network. Due to the fact that the channel in board-to-board communication is deterministic and can be modeled as a strong Line-of-Sight (LoS) channel especially for short range communications. Our new approach considers the optimal antenna arrangement that maximizes the spatial multiplexing gain as shown by works in [3], [4], [5], [6]. For higher frequencies at millimeter waves, the size of the proposed arrays is more compact as compared to designs based on lower frequencies and can be easier embedded on board. Instead of strict parallel boards, the full spatial multiplexing can also be achieved with arrangements on any arbitrary rotated non-parallel planes as indicated by [6], [7], [8], and [9].

In MIMO channels, the transmitted signals are superposed at the receiver side. However, post processing can be applied for parallelization or equalization of the transmitted signals. Works in [10], [11] have shown that analog components like delay lines or phase shifters are capable to create an analog equalizing network that mixes received copies from different antennas and reconstructs the transmitted signals correspondingly. Due to the fact that the sub-channels are deterministic channels, the equalizing networks aiming at creating independent parallel channels have not to be adaptive. By transmitting the signal over the corresponding channel with assistance from a switch network, the computing nodes on the transmitter side can communicate with their targeting receiver as desired. In

Fig. 1: Illustration of the wireless board to board communication with strong LOS MIMO channel. Shown are two computing board with 16 computing nodes each (black). The beams from all transmit antennas (orange) cover the complete receive array. It is aimed to have aligned interference such that a full spatial multiplexing on the links can be achieved.

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case all computing nodes on one board are transmitting to computing nodes on the neighboring board without overlapping of the intended pairs, a full MIMO utilization of strong LoS MIMO channel can be expected. Furthermore, instead of minimizing the undesired interference with pencil beams, received signals with similar magnitudes of their components can be aligned and the undesired interference can cancel each other. Then, the superposed signal can have better SNR and suppresses the interference. Hence, the requirements on the antenna element design, the beamforming capabilities as well as the Butler matrix network can be reduced.

II. OPTIMAL LOS MIMO ARRAY DESIGN

Our previous work [1] has validated the channel model as strong LoS as the power of the second largest channel impulse response is 15 dB less than the direct path. Considering a transmit array ($T_A$) and a receive array ($R_A$) with $N$ antenna elements on both sides, the received signal in a frequency flat strong LoS MIMO channel is modeled as

$$y = \rho(D) \cdot H \cdot s + n, $$

where $s \in \mathbb{C}^{N \times 1}$ is the transmitted signal with unique transmit power $P_T$ at every transmit antenna, i.e., $E(|s_k|^2) \leq P_T$ for $k \in \{1, \ldots, N\}$. Here, we did not follow the approach from E. Telatar [12] that transmit power is shared by the complete array as in practical systems. Instead of that, the transmit power at every transmit antenna is similar even if the antenna number increases. The vector $n$ denotes the white complex Gaussian noise in base-band with $n \sim \mathcal{CN}(0, \sigma_n^2 I_N)$. The magnitude attenuation factor that varies with respect to the transmit distance $D_{lk}$ between $k$-th transmit antenna and $l$-th received antenna is denoted as $\rho(D_{lk})$. As discussed in detail in [13], the attenuation differences on sub-channels can be neglected if the antenna size is much smaller than the transmit distance. For simplicity, we also neglect the differences between the attenuation factors on sub-channels with approximation $\rho(D_{lk}) \approx \rho(D)$, where $D$ is the common distance between the transmit and receive arrays. By $H \in \mathbb{C}^{N \times N}$ we denote the phase coupling matrix in the strong LoS MIMO channel with entities

$$h_{lk} \triangleq e^{-j \frac{2\pi}{\lambda} (D_{lk} - D)}. \quad (2)$$

As indicated by [4], [7], [8], and [14], the optimal antenna arrangements are mainly influenced by the carrier wavelength $\lambda$, the number of antenna elements $N$, and the transmit distance $D$. With high carrier frequency and short distance, the optimal design has the potential for developing ultra-high speed board-to-board communication systems with reasonable antenna sizes. For the parallel uniform linear array (ULA) with $N$ antennas, the optimal antenna spacings $d_t$ and $d_r$ between antennas at transmitter and receiver side given in [7], [8] satisfy

$$d_t \cdot d_r = \frac{\lambda D}{N}. \quad (3)$$

Meanwhile, $d_{t,v}$, $d_{r,v}$, $d_{t,h}$, and $d_{r,h}$ for horizontal and vertical directions of the parallel uniform rectangular array (URA) given in [14] satisfy

$$d_{t,v} \cdot d_{r,v} = \frac{\lambda D}{N_v}, \quad d_{t,h} \cdot d_{r,h} = \frac{\lambda D}{N_h}, \quad (4)$$

where the transceivers consist of $N = N_v \times N_h$ antennas. According to Equation (4), for a $4 \times 4$ MIMO system $(N = 2 \times 2 = 4)$ with symmetric transceivers at high carrier frequencies of 180 GHz and a transmit distance of 10 centimeters, the optimal antenna spacing is given by 9 mm. For simplicity, the system design that is presented in the later sections is based on parallel URA design. However, the system can be adjusted easily via changing the analog equalizing network when non-parallel transceiver boards are considered.

In the latter sections, we investigate fully orthogonal sub-channels $H$ between each antenna pair. With ideal optimal antenna arrangement, the vectors of the phase coupling matrix $H$ are orthogonal to each other with

$$HH^* = N \cdot I_N, \quad (5)$$

where $(\cdot)^*$ denotes the Hermitian transpose operator. However, the validation of this orthogonality is based on an assumption that the pathloss values $\rho(D_{lk})$ on different sub-channels including antenna gains are unique, i.e., $\rho(D_{lk}) = \rho(D)$ for all $l, k \in \{1, \ldots, N\}$ and the transmitted signals are expected to arrive with equal magnitudes. This requires that antenna gains in the direction of corresponding elements should be unique or approximately equal. Therefore, a practical system should be very compact to assure that the geometric and magnitude assumptions hold. For a given number of antennas $N$, the number of antennas in the horizontal and vertical directions $N_h$, $N_v$ of the most compact design follows

$$\arg \min_{N_v \geq N_h} \left( N_v - N_h \right) \quad \text{s.t.} \quad N = N_v \cdot N_h. \quad (6)$$

This array design considers that the uniform rectangular arrays provide less angle spread on the antenna pattern of the elements on the other side compared to ULAs with the same number of antennas, thus being more compact with respect to the antenna aperture.

III. OPTIMAL LOS MIMO SYSTEM DESIGN

A. Analog Signal Equalization

The proposed system is sketched in Fig. 2. The switch network can schedule the transmitted signal from $m$-th computing node from transmitter side ($Tx-m$) to $k$-th transmit antenna ($T_A-k$) in case a communication channel is demanded from $Tx-m$ to $l$-th receiver ($Rx-l$). The switcher can be modeled as a permutation matrix. In this paper, the work is focusing on the maximum transmission rate that can be provided between two boards. Hence, we focus on the analog signal equalization and capacity of the MIMO channel after the switch network.

As a LoS MIMO channel is involved in the system, the received signal $y_l$ at the $l$-th antenna of the received antenna
array is a superposition of all transmitted signals that is modeled as
\[ y_l = \rho(D) \cdot h_l^T \cdot s + n_l, \quad (7) \]
where \( h_l^T = [h_{l1}, h_{l2}, \ldots, h_{lN}] \) is the \( l \)-th row of \( H \) and \( n_l \) is the white complex Gaussian noise with \( n_l \sim \mathcal{CN}(0, \sigma_n^2) \). However, due to the fact that the LoS MIMO channel can be independent sub-channels with a fixed decoupling matrix, the communication system can be employed as parallel SISO channels. Instead of the traditional approach of having the signal processed in digital band which requires high quantization resolution, we propose to utilize an analog equalizing network \( W \) that can decouple the LoS MIMO channel in pass-band. The analog equalizing network can be realized via introducing different delays or fixed phase shifts to the different received signals at different Rx antennas and reconstruct new signals \( y_W \) at the end of the RF chain. We consider that the equalizing network \( W \) is modeled with a fixed realization \( W = H^* \). The \( l \)-th row and \( k \)-th column entry \( W_{lk} \) of \( W \) is suggested to be modeled as a delay line or a phase shifter with \( W_{lk} = h_{lk} \). As described above the equalized signal \( y_W \) is formulated as
\[ y_W = W \cdot y = N \cdot \rho(D) \cdot s + H^* \cdot n, \quad \Delta n \sim \mathcal{CN}(0, N \sigma_n^2 I_N) \] \( \quad (8) \)
where \( \Delta n \sim \mathcal{CN}(0, N \sigma_n^2 I_N) \) is the equivalent Gaussian noise. Note that the analog equalizing network design is based on the Hermitian matrix of the phase coupling matrix.

**B. Link Budget**

When assuming equal power transmission and constant radiated power from each antenna element, the channel capacity of the MIMO transmission given by [15], [12] can be formulated as
\[ C = W \cdot \log_2 \det \left[ I_N + \frac{\rho(D)^2 \cdot P_T}{\sigma_n^2} \cdot HH^* \right] \]
\[ = N \cdot W \cdot \log_2 \left( 1 + \frac{\rho(D)^2 \cdot P_T}{\sigma_n^2} \cdot N \right), \quad (9) \]
where \( W \) is the allocated bandwidth. The noise variance \( \sigma_n^2 \) consists of noise figure \( P_{nf} \) and thermal noise \( P_{th} \). The thermal noise \( P_{th} \) in terms of dBm satisfies
\[ P_{th}[\text{dBm}] = 10 \cdot \log_{10}(1000 \cdot k_B \cdot T \cdot W), \quad (10) \]
where \( k_B \) is the Boltzmann constant (\( k_B = 1.380649 \times 10^{-23} \text{J/K} \)) and \( T \) is the absolute temperature in Kelvin. The power attenuation factor \( P_L(D) = \rho(D)^2 \) in a strong LoS wireless channel can be formulated in [dB] as
\[ P_L(D)[\text{dB}] \triangleq G^{TA} [\text{dBi}] - P_{FE}^{TA} [\text{dB}] - P_{fs}(D)[\text{dB}] + G^{RA} [\text{dBi}] - P_{FE}^{RA} [\text{dB}], \quad (11) \]
where the power of path loss, antenna gain of the Tx and Rx, and the front end loss of Tx and Rx, respectively. The free space path loss \( P_{fs} \) is modeled as
\[ P_{fs}(D) \triangleq \left( \frac{4\pi D}{\lambda} \right)^n_p, \quad (12) \]
where \( n_p \) is the pathloss exponent.

**IV. NUMERICAL RESULTS**

In this section, the capacity of a practical wireless communication MIMO channel between two boards is evaluated numerically. As an example, we consider an LoS MIMO system consisting of 16 antennas on each side with \( N_x = N_y = 4 \) at a transmit distance of 0.1 meter. As suggested by [16], the pathloss exponent \( n_p \) should be 2 and this result is confirmed for short range board-to-board communication by our earlier work [1] in detail. We design the system with a bandwidth of \( W = 30 \text{ GHz} \) and assume that the carrier frequency for board-to-board wireless communication is 180 GHz. The absolute temperature and the noise figure are considered as \( T = 293K \) and \( P_{nf} = 10 \text{ dB} \), respectively.

In our system design, we assume that the transmit signal arrives at different receiver antenna elements with equal magnitude. However, for practical systems the antenna patterns are anisotropic. The spreading angle is less than 16° on the azimuth angle. Therefore, the main lobe of the radiated pattern is recommended to have a beam width larger than 16°. An integrated stacked Vivaldi-Shaped on-chip antenna was designed at 180 GHz in [17]. The taped-out antenna has an antenna pattern with gain of 9.00 dBi in the main direction.
The proposed system in a more realistic system with element gains as shown in Fig. 3. It can be seen that the channel capacities for anisotropic antenna elements are slightly smaller than channel capacities with unique antenna gains. However, the proposed channel model is a good approximation to the realistic channel with anisotropic antenna patterns.

V. CONCLUSIONS

The proposed new computer design relies on wireless interconnects based on the so called Line-of-Sight MIMO approach. Instead of interference suppression which comes with the burden of large antenna arrays and lossy feeding networks, the proposed approach exploits the interference in terms of analog equalization. By considering real antenna measurement data of a 180GHz Vivaldi antenna and optimal
antenna positioning it has been shown that the corresponding channel is nearly unitary which allows for delay based equalization in analog domain. This enables full spatial multiplexing gains which can be utilized in a flexible fashion like parallel energy-efficient independent SISO channels. The numerical evaluation shows that 1 Tbps and above can be achieved at moderate transmit power which obviates the need for power amplifiers.

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REFERENCES


