2014 IEEE International Conference on Acoustic, Speech and Signal Processing (ICASSP)

FEASIBILITY STUDY ON FULL-DUPLEX WIRELESS MILLIMETER-WAVE SYSTEMS

Liangbin Li⋆ *Kaushik Josiam* † *Rakesh Taori*†

⋆ University of California, Irvine, CA
† Samsung Research America, Dallas, TX

ABSTRACT

To enable simultaneous transmit and receive operations on the same frequency band, also known as full duplex, on adjacent panels of a millimeter wave (mmWave) base station (BS), we benchmark the amount of leakage interference from the transmitter to the receiver. We construct a *self-interference channel* model consisting of line-of-sight (LOS) leakage and non-LOS (NLOS) reflections from nearby clusters. From simulations, we further show that due to antenna beamforming in mmWave systems, the LOS leakage interference is subdued, whereas the NLOS leakage is amplified. For some BS designs, the NLOS leakage is likely to be the dominant interference. It impacts system designs in two ways. First, the worst case interference can still saturate the receiver chain and would require analog and digital cancellation in addition to the isolation provided by antenna beamforming. Second, due to the delay associated with NLOS reflective components, the cancelers would need to accommodate a large dynamic delay.

1. INTRODUCTION

Wireless communication systems have traditionally considered the use of separate transmit and receive processing chains connected to a common antenna via a duplexer. The resulting interference from the transmit chain into the receive chain necessitates the separation of transmit and receive operations in either time or frequency. When systems operate by division in time or frequency, they are said to be in half duplex mode where improved signal strength comes at the expense of throughput. Even when transceiver chains are connected to different antenna, the proximity of the antennas causes enough transmission leakage into the receiver chain to make full duplex difficult. For such systems, a combination of passive antenna arrangement and active analog/digital cancellation techniques [1, 2] has been used to realize full duplex operation in the microwave ($<$ 3GHz) band. In a recent effort [3], significant advances have been reported for full duplex operation in a traditional single antenna system by using a combination of active analog and digital cancelers that reduce leakage interference to the levels below the noise floor. So far, research on full Duplex has primarily focused on operation in microwave frequencies. In this paper, we will investigate the feasibility of full duplex at millimeter-Wave (mmWave) frequencies.

This work was done during the first author's summer internship with Samsung Research America, Dallas.

Signals modulated by mmWave carriers experience larger free space path loss and undergo significant reflection losses [4]. To overcome these effects, beamforming in a desired direction is performed using high-gain antenna arrays. Beamforming involves exciting each antenna in the array by phase shifted versions of a signal. The phase shifted signals from the antennas form a beam by adding constructively in a desired direction and destructively in others making the radiated power an angular function. The beams can be steered by altering the phase shifts on the signal.

To study feasibility of full-duplex systems at mmWave, we consider transmit and receive chains connected to two different linear arrays. The transmit and receive arrays can be arranged either on a single plane or on two different planes. In both arrangements, leakage from the transmit array can overwhelm the received signal at the receive array in full duplex operation. In this paper, we consider the two plane arrangement of the arrays shown in Fig.1. The setting can be considered for applications involving mmWave backhaul. The contributions of this paper are two-fold:

- We build a model for the leakage channel that consists of Line-of-sight (LOS) and reflective non-LOS (NLOS) components. We show empirically that the dominant leakage, *when beamforming is used*, could be from the NLOS component. This is in contrast to the observations made in studies on microwave band full duplex, where interference is dominated by the LOS channel [5].

- We characterize the distribution of the leakage power from the transmitter as seen by a beamformed receiver and quantify the amount of cancellation necessary to reduce the interference. In doing so, we are able to identify the requirements on the analog cancellation to reduce the interference to the levels that are within the dynamic range supported by the analog to digital converter (ADC), after which a digital canceler can be employed to further suppress the interference.

The rest of the paper is organized as follows. We describe the self-interference channel in Section 2. The simulation results are shown in Section 3 followed by conclusions in Section 4.

978-1-4799-2893-4/14/$31.00 ©2014 IEEE

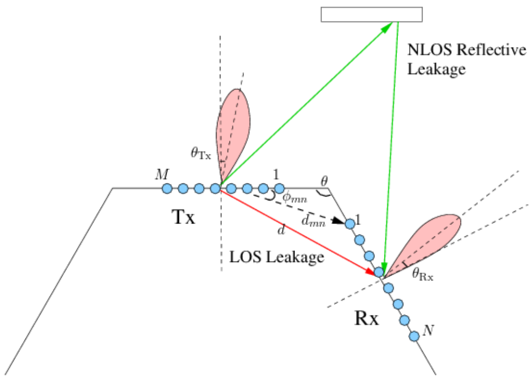

Fig. 1. Full-duplex transmission on adjacent panels of a six-panel base station. Linear antenna arrays are deployed at both transmitter and receiver. There are two types of leakage: LOS leakage (red color) and reflection leakage (green color).

2. THE SELF-INTERFERENCE CHANNEL

Consider the transceiver illustrated in Fig. 1. We denote the angle between two adjacent panels as θ. For six-panel BS, $\theta = \frac{2\pi}{3}$. The center of the two arrays are separated by a distance of d meters. The transmitter has an array of M antenna elements, whereas the receiver has an array of N antenna elements. The antenna spacings are assumed to be equal to half of the wavelength. The transmit and receive antenna arrays can be steered towards a spatial direction with maximum antenna gain by choosing an appropriate phase shift at each antenna. Let us assume that the steered angles are θ_{Tx} and θ_{Rx} for transmitter and receiver, respectively, measured w. r. t. the bore-sight of each array. For simplicity, we present the channel model for one-dimensional linear arrays. Extension to two-dimensional arrays is straightforward.

The self-interference channel is made up of two components. The first component is a LOS path through the back side of linear array (represented by red color in Fig. 1). The second component is a multi-path NLOS leakage due to reflections from nearby clusters (represented by green color in Fig. 1). Transmit signal from the array is reflected (or diffracted) by nearby clusters, and arrives at the receive array.

2.1. The LOS path

Typically, the signals from the transmitter are assumed to impinge on a receive array in the far field as a planar wave. The planar wave assumption results in two approximations. First, all antenna elements observe a signal with the same path loss. Second, the phase difference among antenna elements depends only on the angle of arrival (AoA) and the receive antenna spacing. The planar wave assumption does not hold when the transmit and receive arrays are on two close-by adjacent panels. Consider the indexing of antennas as shown in Fig. 1. Due to the proximity of two arrays, the AoA at receive Antenna 1 from transmit Antenna 1 is different compared to that from transmit Antenna 8. The difference is enlarged when the panel angle θ is decreased. Therefore, we consider a spherical propagation model for the LOS leakage path. Denote the ray between transmit Antenna m and receive Antenna n as Ray (m,n). The distance between transmit Antenna m and receive Antenna n is denoted as d_{mn}. Clearly, the arrival phase and path loss correspond to Ray (m,n) depend on the distance d_{mn}. We model the LOS path of Ray (m,n) as $h^{\mathrm{LOS}}_{m,n} = \sqrt{\beta_{mn}}\exp\left(-j\frac{2\pi}{\lambda}d_{mn}\right)$, where λ and $\beta_{mn} = \left(\frac{\lambda}{4\pi d_{mn}}\right)^2$ denote the wavelength and the free-space path loss for the ray, respectively. Denote the angle between transmit panel and Ray (m,n) as ϕ_{mn}, and that between receive panel and Ray (m,n) as ψ_{mn}. From triangle geometries, we have $\phi_{mn} + \psi_{mn} + \theta = \pi$. Let the antenna gain of each transmit and receive element at the direction of Ray (m,n) be $G_{\mathrm{Tx}}(\phi_{mn})$ and $G_{\mathrm{Rx}}(\psi_{mn})$, respectively. Further denote the steering phase shifts as a_m for transmit Antenna m and b_n for receive Antenna n. The equivalent channel at the output of receive Antenna n can be written as

$$\bar{h}^{\mathrm{LOS}}_n = \sum_{m=1:M} \sqrt{G_{\mathrm{Tx}}(\phi_{mn})G_{\mathrm{Rx}}(\psi_{mn})}\, a_m b_n h^{\mathrm{LOS}}_{m,n}. \quad (1)$$

To steer the transmit array towards the desired direction θ_{Tx}, we have $a_m = \frac{1}{\sqrt{M}}\exp((m-1)\pi j \sin\theta_{\mathrm{Tx}})$. Similarly, we have $b_n = \frac{1}{\sqrt{N}}\exp((n-1)\pi j\sin\theta_{\mathrm{Rx}})$ for receive array towards direction θ_{Rx}. Combining signals from all receive antenna ports, we have an equivalent channel for LOS path as

$$h^{\mathrm{LOS}} = \sum_{m=1:M,n=1:N} \sqrt{\beta_{mn} G_{\mathrm{Tx}}(\phi_{mn}) G_{\mathrm{Rx}}(\psi_{mn})} \times a_m b_n \exp\left(-j\frac{2\pi}{\lambda}d_{mn}\right). \quad (2)$$

When d is large, the above model can be approximated based on planar wave by assuming $\beta_{mn} \approx \beta$, $\phi_{mn} \approx \phi$, $\psi_{mn} \approx \psi$, and $d_{mn} \approx d + (m-1)\cos(\phi)\frac{\lambda}{2} + (n-1)\cos(\psi)\frac{\lambda}{2}$ for $\forall m, n$.

2.2. The reflective path

The reflective path is due to reflection or diffraction from nearby clusters. We assume a spatial channel model [6] made up of L clusters, each with delay, τ_l , $l = 1, \ldots, L$. Each cluster further consists of Q rays in different spatial angles. We associate each ray with an angle-of-departure (AoD) ϕ^{Ref}_{lq}, $l = 1, \ldots, L, q = 1, \ldots, Q$, at the transmit array and an AoA ψ^{Ref}_{lq} at the receive array. Since the propagation distance of the reflective paths is larger than that of the LOS path, planar wave assumption holds for all reflective paths. We can compute the overall array gain at direction ϕ^{Ref}_{lq} when the array is steered towards θ_{Tx} as

$$A(\theta_{\mathrm{Tx}}, \phi^{\mathrm{Ref}}_{lq}, M) = G_{\mathrm{Tx}}(\phi^{\mathrm{Ref}}_{lq}) \times \left| \sum_{m=1:M} a_m \exp\left(-(m-1)\pi j(\sin\phi^{\mathrm{Ref}}_{lq})\right)\right|^2, \quad (3)$$

where the steering coefficient a_m is a function of θ_{Tx}. The first term is the gain offered by each antenna element, whereas the second term is the linear array gain. Similarly at the receive array, we can compute the overall array gain function as $A(\theta_{\mathrm{Rx}}, \psi_{lq}^{\mathrm{Ref}}, N)$. At the output of receive array, we model the channel gain associated with Cluster l and ray q as

$$h_{lq}^{\mathrm{Ref}} = \sqrt{\beta_l A(\theta_{\mathrm{Tx}}, \phi_{lq}^{\mathrm{Ref}}, M) A(\theta_{\mathrm{Rx}}, \psi_{lq}^{\mathrm{Ref}}, N) p_{lq}} \exp(j\Phi_{lq}), \tag{4}$$

where β_l, p_{lq}, and Φ_{lq} denote the path loss of all rays from Cluster l, the normalized power of the ray associated with Cluster l and Ray q, and phase transition due to reflection, respectively. To normalize power, it is required that $\sum\limits_{l=1:L,q=1:Q} p_{lq} = 1$. Without loss of generality, we can assume that clusters are indexed based on their distance to the transceiver. Then, Cluster l has the lth smallest delay, i.e., $\tau_1 \leq \tau_2 \leq \ldots \leq \tau_L$. To model the path loss of Cluster l, we assume clusters are randomly dropped with a distance to the transceiver uniformly distributed in $[d_{\min}, d_{\max}]$. The resulting L distances are sorted in an ascending order. The l-th distance after sort, denoted as d_l^{Ref}, is associated with Cluster l. Then, β_l is modeled as

$$-10\log_{10}\beta_l = \beta_0 + \alpha \times 10\log(2d_l^{\mathrm{Ref}}) + \xi, \tag{5}$$

where α denotes the path loss exponent, and ξ denotes reflection loss. We apply the best NLOS path loss exponent, $\alpha = 4.56$, reported in [4] measured using a reference distance of 5m, i.e., the path loss equation is $10\log_{10}\left(\frac{4\pi\times 5}{\lambda}\right)^2 + 10\alpha \times \log_{10}\left(\frac{d}{5}\right)$. Comparing the path loss equation with (5), gives $\beta_0 = -20\log_{10}\left(\frac{\lambda}{4\pi\times 5}\right) - 10\alpha\log 5 = 43.5$ dB. We assume that the reflection loss is $\xi = 10$ dB.

The self-interference channel consists of both LOS leakage and reflective leakage. Assume that the LOS path has zero delay. Combing Eq. (2) with (4), the time-domain self-interference channel can be expressed as

$$h(t) = h^{\mathrm{LOS}}\delta(t) + \sum_{l=1:L,q=1:Q} h_{lq}^{\mathrm{Ref}}\delta(t-\tau_l), \tag{6}$$

where $\delta(t)$ denotes the delta function.

3. TIME-DOMAIN CHARACTERISTICS

Eq. (6) defines a time-domain multipath channel after antenna beamforming. In this section, we study the temporal characteristics of the channel and simulate the impact of LOS and reflective leakages provided by the proposed model. We first consider a transceiver design with six panels where the angle between adjacent panels is $\theta = \frac{2\pi}{3}$. The array separation distance is $d = 0.5$ meters. Other transceiver designs will be compared later. We consider a linear antenna array that has 8 antenna elements whose beam pattern at $\theta_{\mathrm{Tx}} = 0$ is shown in Fig. 2. For the reflective paths, the AoD ϕ_{lq}^{Ref}, AoA ψ_{lq}^{Ref},

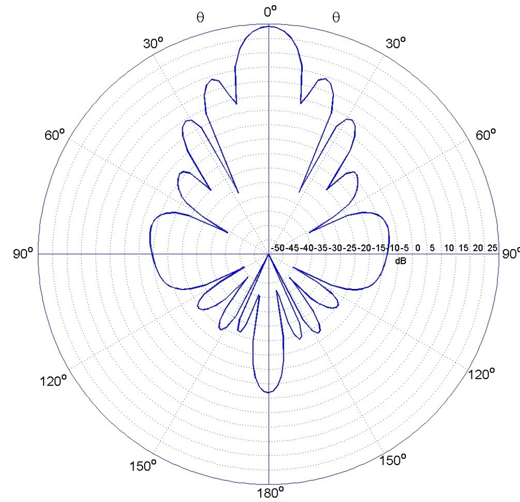

Fig. 2. The overall array gain (beam pattern) $A(0, \theta, 8)$ with the steering angle $\theta_{\mathrm{Tx}} = 0$. At the boresight, the antenna gain is approximately 24 dB.

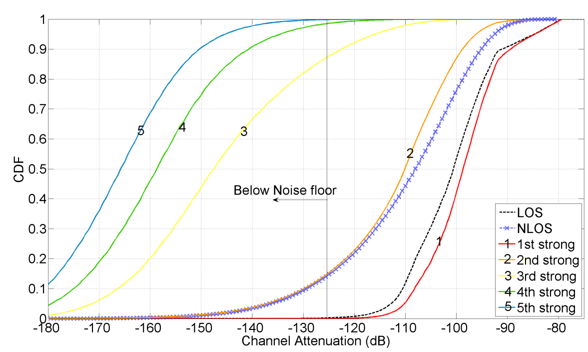

Fig. 3. CDF of channel attenuation for different paths.

delay τ_l, phase transition Φ_{lq}, and the normalized power p_{lq} are generated according to the 19-ray urban micro environment in spatial channel model [6]. We further assume that the distance between clusters and the transceiver is uniformly distributed in $\mathcal{U}[20, 50]$ meters. The steering angles $\theta_{\mathrm{Tx}}, \theta_{\mathrm{Rx}}$ are uniformly distributed in $\left[-\frac{\pi}{6}, \frac{\pi}{6}\right]$ since they will be steered in the directions of their desired links.

The cumulative distribution function (CDF) of channel attenuation and path delay, with 1 GHz bandwidth at 28 GHz frequency, are shown in Figs. 3 and 4, respectively. The curves labeled 'LOS' and 'NLOS' correspond to the attenuation of the LOS path and the total power of all reflective paths, respectively. The worst-case isolation of LOS path is around 80 dB, whereas that of all reflective paths is around 85 dB. We also show the CDF of the ordered gain of the first five strongest paths (labeled 'nth strong'). Note that for a mmWave system with 1 GHz bandwidth, the noise floor is around -80 dBm. Let the transmit power be 46 dBm. Any path with gain less than -126 dB would be below the noise floor, and it is indistinguishable from noise by the receiver.

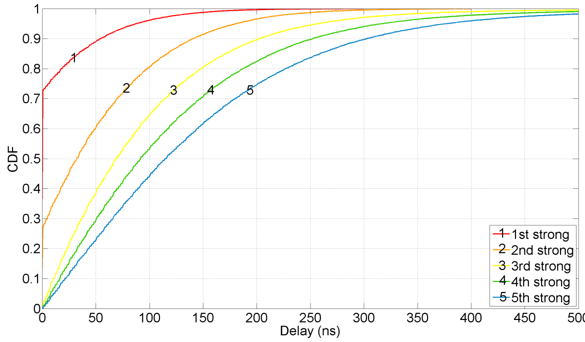

Fig. 4. CDF of channel delays for different paths.

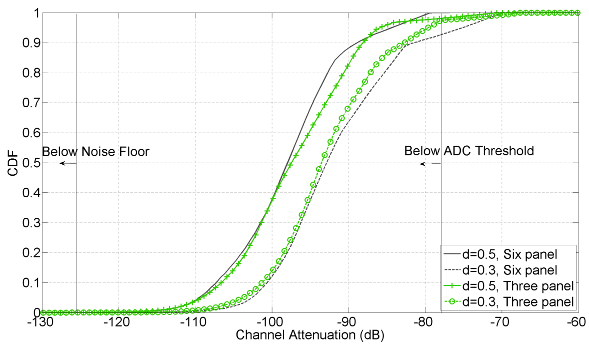

Fig. 5. CDF of total channel attenuation for different transceiver designs.

Therefore, the effective number of the multipath is four for the case in Fig. 3. Fig. 4 shows the CDF of delays of the first five strongest paths. Since the LOS path has 0 delay and any nonzero delay is contributed by the reflective path, we can observe from Fig. 4 that the strongest channel are 73% of the time from the LOS path and for the remaining 27% of the time, it is from the reflective paths. Recall that on microwave frequency, full-duplex system is without antenna beamforming, thus the dominant interference is from LOS leakage. On mmWave frequency, the dominant path is likely to be from reflection of nearby clusters due to antenna beamforming.

We further provide an example to show how antenna arrays reshape the leakage power distribution between LOS and reflective paths. Let us assume the steering angles $\theta_{\rm Tx}, \theta_{\rm Rx}$ equal to zero and one cluster is 20 meters away from the transceiver. Also let the AoD $\phi_{lq}^{\rm Ref}$ and AoA $\psi_{lq}^{\rm Ref}$ be 10 degree. We summarize the path-loss and the equivalent path-loss after antenna beamforming for LOS and reflective paths in Table. 1. Note that the equivalent path loss of the reflective path can be obtained by adding antenna array gain to the path loss. From Fig. 2, the antenna array gain at AoD$= 10$ degree is approximately 15 dB. The total transmit and receive antenna gain is 30 dB. It can be observed that at the output of antenna port, the equivalent path loss of the LOS path is enlarged because the antenna arrays provide additional isolation. However, the equivalent path loss of the reflective path is decreased since the antennas amplify the front-end leakage. Consequently, the reflective leakage is stronger than the LOS leakage after antenna beamforming for this example.

Table 1. Link budget analysis for LOS and reflective paths at $\theta_{\rm Tx}, \theta_{\rm Rx} = 0$ and $d_l^{\rm Ref} = 20$ meters . AoD, AoA=10 degrees.

	LOS Path	Reflective Path
Path Loss	-55.3 dB	-126.5 dB
After Antenna Array	-98.2 dB	-96.5 dB

Fig. 5 shows the CDF of total power of all paths (including the LOS path and all reflective paths) for different transceiver designs. Let us assume that the receiver uses an ADC with 10 effective bits. If the receiver allows the self-interference to access at most 8 effective bits, then the tolerable interference-to-noise ratio is about 48 dB above the noise floor. Thus, with a noise floor corresponding to channel gains at -126 dB in Fig. 5, the ADC threshold will be at $-126{+}48 = -78$ dB. The worst-case channel attenuation for a six panel transceiver with $d = 0.5$ meters is -80 dB, and does not saturate the receiver's ADC threshold at -78 dB. Thus, for this specific transceiver design, we need baseband algorithms to cancel the $126{-}80 = 46$ dB residue leakage interference. For the other three transceiver designs, the worst-case interference leakage is above the ADC threshold. We need to either improve the ADC threshold or design cancellation algorithms in analog domain. Generally, adding ADC effective bits consumes more power and requires challenging hardware designs [3]. Analog cancellation algorithms, e.g., [3], can be used to mitigate leakage interference, if the algorithm can provide a large dynamic range to track the delay of the peak leakage path. The required dynamic range for the delay path is from 0 to 200 ns, as observed from Fig. 4.

4. CONCLUSIONS

We proposed a methodology to model the leakage channel from the transmit to the receive array in a full duplex mmWave transceiver. Using the self interference channel model, we showed that beamforming reduces the impact of the LOS component. At the same time, the received power from the NLOS reflective components increase and contribute to a major portion of interference seen at the receiver. The distribution of the interference power shows a large spread in its values where the worst case is about 46 dB above the noise floor, implying that we cannot only rely on beamforming to provide isolation between transmit and receive arrays. Instead, both analog and digital cancellation are important to realize the promise of mmWave full duplex systems. Future work includes modeling impairments due to components in the transceiver chain and proposing analog/digital cancellation algorithms to suppress about 46 dB of interference.

5. REFERENCES

[1] M. Duarte, C. Dick, and A. Sabharwal, "Experiment-driven characterization of full-duplex wireless systems," *IEEE Transactions on Wireless Communications*, vol. 11, no. 12, pp. 4296–4307, 2012.

[2] J. Choi, M. Jain, K. Srinivasan, P. Levis, and S. Katti, "Achieving single channel, full duplex wireless communication," 2010, MobiCom, pp. 1–12.

[3] D. Bharadia, E. McMilin, and S. Katti, "Full duplex radios," in *Sigcomm*, 2013.

[4] T.S. Rappaport, S. Sun, R. Mayzus, H. Zhao, Y. Azar, K. Wang, G.N. Wong, J.K. Schulz, M. Samimi, and F. Gutierrez, "Millimeter wave mobile communications for 5G cellular: It will work!," *IEEE Access*, vol. 1, pp. 335–349, 2013.

[5] E. Everett, A. Sahai, and A. Sabharwal, "Passive self-interference suppression for full-duplex infrastructure nodes," Jan. 2013, [Online]. Available: http://arxiv.org/abs/1302.2185.

[6] ITU-R, "M.2135-1 Guidelines for evaluation of radio interface technologies for IMT-Advanced," Tech. Rep., Dec 2009.