Frequency-Selective Analog Beam Probing for Millimeter Wave Communication Systems

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Abstract—This work focuses on the initial beam acquisition/alignment of millimeter wave (mmWave) communication systems. To detect the angle of arrival (AoA) and/or angle of departure (AoD), we propose a training protocol which probes all beamformers from a given codebook simultaneously by exploiting the sparse nature of mmWave channels. By applying a frequency-selective beam probing network, we can map each beamformer from the codebook to different frequencies and a spectral analysis at the receiver allows us to deduce favorable beamformers or AoDs. For practical reasons, we elaborate this idea of steering direction to frequency mapping for an orthogonal frequency division multiplexing (OFDM) communication system, i.e., we map each beamformer to specific pilot subcarriers. Under two different hardware designs, we investigate the feasibility of building such beamformer to frequency mappings for one additional radio frequency (RF) chain next to an existing OFDM communication system. We show that parallel beam training is able to achieve better effective transmission rates than exhaustive search in fast-time varying environments due to high temporal efficiency. This is crucial for mmWave communication systems which have access to large beamforming codebooks but suffer from short coherence times due to mobility and high spatial resolution.

Index Terms—Beam Acquisition, Beam Alignment, Orthogonal Frequency Division Multiplexing, Spatial Wideband Effect, Beam Squint

I. INTRODUCTION

Enabling millimeter wave (mmWave) communication is key to accommodate demand for gigabit data transmission in future mobile systems as large spectrum of unlicensed bandwidth is available in this regime. However, carrier frequencies $f_c$ in the millimeter wave range come at a cost of higher propagation losses which need to be mitigated [1]. By employing directional antenna arrays at the transmitter (Tx) and receiver (Rx), a set of narrow beams with high beamforming gains can compensate the aforementioned high path loss [2], [3]. Consequently, we need to account for additional training overhead in form of initial beam acquisition/alignment before data transmission [4].

During beam alignment, Rx and Tx try to detect their angle of arrival (AoA) and angle of departure (AoD) to maximize either the detection probability or transmission rate at the core [5]. Hereby, the spatial domain can be probed either using downlink or uplink. In this work, we will focus on downlink beam probing meaning that the base station periodically starts a directional search to discover user equipments within cell coverage.

Most commonly this directional search is performed by the base stations by randomly selecting from a given beamforming set/codebook various beamformers to test iteratively different spatial angles. This approach, also called exhaustive search, is optimal in the sense that we test all available beamformers over the complete bandwidth for different time instant $n$. A schematic of this procedure is drawn in Fig. 1, where we further assume an orthogonal frequency division multiplexing (OFDM) symbol with certain subcarriers, e.g., $f_0 \leq f_1, \ldots, \leq f_3$, that is fed to the input ports of a beamforming network. Selecting different input ports over time instant $n$, a beamforming network, e.g., Butler matrix, is capable of producing and testing distinct AoDs. After the minimum amount of $|\mathcal{F}_P|$ steps, where $\mathcal{F}_P$ is the beamforming codebook, the Rx has “seen” all available beamformers and is able to infer the best beamforming vector. With this the minimum number of time instants, i.e., the timing overhead, is limited by the size of the beamforming codebook $\mathcal{F}_P$. However, this timing overhead is an unrealistic lower bound as in practical hardware realizations we are usually not able to switch between several
beamformers from time instant to time instant. In literature [6], we can find the idea of hierarchical search to speedup the initial beam alignment process. However, the major drawback of hierarchical search is often overseen as the wide beams in the first stages can produce only small beamforming gains. Incorrect decisions in the first steps can easily produce training misalignment errors.

In contrast, we propose in this work a beam probing procedure which does not rely on fast switching between various beamformers and always uses beamformers with maximum beamforming gain. Furthermore, as we test all available beamformers from codebook \( \mathcal{F}_p \) simultaneously, the training overhead is minimized. Our approach is illustrated in Fig. 2, where one can derive that we use distinct frequencies/subcarriers \( f_0, \ldots, f_3 \) and map them to distinct beamformers. This can be done by feeding individual frequency/subcarriers into separate input ports of a beamforming network, e.g., Butler matrix. However, we will show later how a beamforming network with delay elements can be implemented which is capable of producing all beamformers from \( \mathcal{F}_p \) in a novel and more flexible way. A single antenna receiver in the far field can perform a spectrum analysis and as each AoD or spatial frequency is mapped to a certain temporal frequency, the Rx can deduce its best beamformer. With this approach, we are able to obtain preliminary estimates for each beamforming vector from \( \mathcal{F}_p \) at the very first time instant, i.e., \( n = 1 \).

In this work, we will investigate designs and algorithms for parallel beam probing. We will derive that especially for large codebook sizes and few scatterers the timing overhead can be dramatically reduced as we prevent spending time resources on non-promising beams. As a consequence, the reduction of the timing overheads for the initial beam acquisition phase enables mmWave communication for scenarios which are dominated by short coherence times, like vehicle-to-vehicle (V2V) communication or other applications in fast-time varying environments.

II. SYSTEM MODEL

The investigation to speedup the beam alignment phase is limited by the sparsity and the frequency-selectivity of the channel. Therefore, we examine a mmWave geometrical channel model, i.e., only a few scatterers in the environment, which is widely used in literature [6]. Further, we assume OFDM for practical relevance as this allows, in our opinion, a better integration of our parallel beam probing network into an existing high-frequency (HF) data transmission component. Without loss of generality, our idea will be presented for a Tx equipped with a uniform linear array (ULA) with \( M_{Tx} \) antenna elements and an Rx with only one receive antenna. The environment is modeled under the assumption of several paths between the Tx and Rx, where each path \( l \) is fully described by its complex Gaussian gain \( a_l \), delay \( \tau_{l,m}^{ch} \) and AoD \( \phi_{Tx,l} \). Typically, we assume that these parameters stay constant for various time instants and we can define a block of size \( N_{coh} \) of channel uses with constant gains, delays, and AoDs realizations. Such a blockfading channel model is commonly used in mmWave literature and follows the geometrical channel model assumption in [6]. With this in mind, we define the channel impulse response (CIR) between the \( m \in \{0, \ldots, M_{Tx} - 1\} \)-th transmit antenna and the receiver of our multiple-input single-output (MISO) system in baseband as [7]

\[
c_m(t) = L \sum_{l=1}^{L} \alpha_l e^{-j2\pi f_m \tau_{l,m}^{ch}} \delta(t - \tau_{l,m}^{ch}), \quad (1)
\]

where the progressively added delay \( \tau_{l,m}^{ch} \) of path \( l \) from the \( m \)-th Tx antenna to the Rx antenna is found by

\[
\tau_{l,m}^{ch} = \tau_{l}^{ch} + m \frac{d \sin \phi_{Tx,l}}{c}.
\]

Hereby, \( L \) is the total number of paths or scatterers and \( f_c \) denotes the carrier frequency of the transmitted signal. Variable \( d \) denotes the antenna element spacing and \( c \) represents the speed of light. This follows the line of argument in [8] and models a wideband mmWave channel model in the time domain for \( t < T_{coh} \), where \( T_{coh} \) is the coherence time of the channel. In (2), we model additionally to the timing delay \( \tau_{l}^{ch} \) of the path \( l \), a timing offset based on the array geometry which cannot be neglected if the array size becomes large or the transmission bandwidth \( B \) increases. Taking the Fourier transform of (1), the channel transfer function in baseband, \( 0 \leq f \leq B \), can be written as

\[
h_m(f) = L \sum_{l=1}^{L} \alpha_l e^{-j2\pi f_m \tau_{l,m}^{ch}} e^{-j2\pi \tau_{l,m}^{ch} f}.
\]

By stacking all \( h_m(f) \), we obtain a vector-sized channel transfer function \( \mathbf{h}(f) \) which can be simplify to

\[
\mathbf{h}(f) = \sum_{l=1}^{L} \alpha_l \mathbf{a}_f^H(\phi_{Tx,l}) e^{-j2\pi \tau_{l}^{ch}(f + f_c)}
\]

by applying (2) to (3) and defining

\[
\mathbf{a}_f(\phi_{Tx,l}) = \left[ 1, \ldots, e^{j2\pi (M_{Tx}-1) \frac{\sin \phi_{Tx,l}}{\lambda (1+f_c)}} \right]^T
\]

as the frequency dependent array response vector [8].

Next, for an OFDM system with \( K \) subcarriers we divide the MISO channel from (4) into \( K \) flat-fading subchannels in the frequency domain, i.e., we assume \( K \) parallel complex additive white Gaussian noise (AWGN) channels. With OFDM symbol rate \( f_s = B/K \), the value of the channel transfer function at the \( k \)-th subcarrier, \( k \in K = \{0, \ldots, K-1\} \), can be obtained by sampling (4)

\[
\mathbf{h}_k \equiv \mathbf{h}(f = k f_s) = \sum_{l=1}^{L} \bar{\alpha}_l \mathbf{a}_{f_s}^H(\phi_{Tx,l}) e^{-j2\pi \tau_{l}^{ch} f_s k},
\]

where \( \bar{\alpha}_l = \alpha_l e^{-j2\pi \tau_{l}^{ch} f_s} \). The set of \( \mathbf{h}_k \) for \( k \in \{0, \ldots, K-1\} \) describes the interaction of the ULA at the transmitter with the random channel environment.
which are fed into two distinct

(a) Schematic of the filtering beam probing network with fixed active subcarrier set of size $|\mathcal{K}| = 8$

(b) Schematic of the frequency-selective beam probing network with active subcarrier set of size $|\mathcal{K}| \geq 8$

Fig. 3: HF parallel beam probing networks for a ULA with $M_{\text{Tx}} = 8$ using the typical OFDM transmitter chain

III. HARDWARE IMPLEMENTATION

In the following section, we describe two possible hardware implementations (see Fig. 3) for parallel beam probing. In both cases, the probing subcarriers $k$ are defined in frequency-domain which can be converted into time-domain after serial-to-parallel (S/P) conversion and taking the Inverse Fast-Fourier-Transform (IFFT). The parallel-to-serial (P/S) converter together with the RF-Chain produce a single analog transmit signal at an intermediate or carrier frequency $f_c$ which are fed into two distinct $1 : M_{\text{Tx}}$ networks. As RF chains in communication systems are quiet costly and consume a lot energy, we restrict ourselves to using only one RF chain for such an additional HF beam probing component. To the best of our knowledge, we propose with the latter design a novel idea to improve the beam probing for mmWave communication.

A. Filtering Beam Probing Network

In the first design, a set of known active pilot subcarriers $\mathcal{K}$ of size $M_{\text{Tx}}$ produces the probing OFDM symbol as shown in Fig. 3a. Such an OFDM symbol can be processed by a filtering network, where a combination of matched lowpass and highpass filters maps over $\log_2 |\mathcal{K}|$ stages each active subcarrier to one output port. These filtered signals (Inp. Bf.) can be multiplexed into inputs of an existing data transmission HF beamforming network such that we can guarantee to perform the beam probing on the same hardware (see Fig. 2). Alternatively, a band of bandpass filters can be used to obtain the active subcarriers at the outputs of a $1 : M_{\text{Tx}}$ network.

B. Frequency-Selective Beam Probing Network

In contrast, using the network from Fig. 3b and producing all the beamformers with the frequency-selective beam probing network, we need to take into account that during beam probing a different hardware is used then later during data transmission. However, we can perform a more flexible mapping of subcarriers/frequencies to beamformers with such a beamforming network. Delaying antenna input ports (Inp. Ant.) by multiples of $\tau_d$ will lead to various progressive phase shifts between neighboring antenna elements per subcarrier/frequency. With this, a beam probing codebook in the discrete case of size $|\mathcal{F}_p| = |\mathcal{K}| \geq M_{\text{Tx}}$ is obtained and also a subset of size $|\mathcal{K}| = M_{\text{Tx}}$ mimicking the classical orthogonal codebook $\mathcal{F}_p^\perp$ can be constructed. Additionally, we would like to refer the reader to [9], where we present the beneficial behaviour of using all subcarriers and obtaining an over-sampled channel transfer function at the receiver. Summing up, our concept turns the negative wideband effect or beam squint into a positive feature where we can test various AoDs in parallel.

To understand this approach, we need to derive the frequency dependent beamforming vector of the delay network in Fig. 3b

$$f_{p,k} = \frac{1}{M_{\text{Tx}}} \left[ e^{-j2\pi f_c \tau_0 (1 + \frac{k}{K})}, ..., e^{-j2\pi f_c \tau_{M_{\text{Tx}}-1} (1 + \frac{k}{K})} \right]^T$$

(7)

where $\tau_0 = i \tau_d = i \frac{M_{\text{Tx}}-1}{M_{\text{Tx}}}$, $i \in [0, ..., M_{\text{Tx}}-1]$ are the fixed timing delays. By sampling (7) according to the OFDM symbol rate $f_s$, we get a discrete set of beamformers $\{f_{p,k}\}$ for $k \in 0, ..., K-1$, or a subcarrier based beam probing codebook, respectively. Using only $\mathcal{K} \subset \mathcal{K}$ as active subcarriers, where $\mathcal{K} = \{0, ..., M_{\text{Tx}}-1\} \setminus \{K / (M_{\text{Tx}}-1)\}$, the corresponding beam probing codebook $\{f_{p,k}\}$ is equivalent to a two-dimensional Discrete Fourier Transform matrix and the classical orthogonal codebook $\mathcal{F}_p^\perp$ is obtained for this specific subcarrier set. Important to note is that the progressively added timing delay $\tau_d$ between neighboring antennas is based on the given bandwidth $B$ and $M_{\text{Tx}}$ such that the beam probing network can be designed for any intermediate or carrier frequency. The tree structure shown in Fig. 3b also helps hardware designers to reuse certain building blocks and simplifies designing the frequency-selective beam probing network.

As (5) and (7) are function of baseband frequency $f$, the inner product of both, which is commonly known as the
Assuming CSI at the receiver and (9), the Rx needs to solve

\[ A(\phi, f) = a_{f, T_Tx}^H(\phi, f) f_{p, f} \]

\[ = \frac{1}{M_{\text{tx}}} \sum_{m=0}^{M_{\text{tx}}-1} e^{-j2\pi m \left( \frac{\tau_d}{M_{\text{tx}}} - \frac{\eta}{4} \sin(\phi) \right) \left( 1 + \frac{f}{f_c} \right)}, \]

where \( \tau_d = \frac{\tau}{M_{\text{tx}}} \) and \( \eta = B/f_c \) the relative bandwidth. For a typical mmWave system, \( f_c = 60\text{GHz}, B = 2\text{GHz}, M_{\text{tx}} = 8, \tau_d \approx 0.44\text{ns}, \mathcal{K} = \{0, 292, 584, 876, 1168, 1460, 1752, 2044\} \), the sampled \( A(\phi, f) \) is shown in Fig. 4b. In Fig. 4a, we can see that the spatial signature of the subcarrier based beam probing codebook of the frequency-selective beamformer mimics the classical wideband orthogonal codebook \( \mathcal{F}_p^\perp \). By choosing additional and/or different subcarriers \( k \not\in \mathcal{K} \), we can refine and/or change the resolution of the AoD detection [9]. Changing the bandwidth and/or carrier frequency and using different relative bandwidths \( \eta \) will lead to a cyclic shift of the mapping between steering direction and frequency [10].

IV. ALGORITHMS FOR INITIAL BEAM ACQUISITION

In this section, we would like to focus on two signal processing algorithms, one for the exhaustive search and one for the parallel search. During exhaustive search, we test each analog beamformer time after time and, therefore, classify it as iterative beam probing. However, our proposal test beamformers parallel in time and we classify it as parallel beam probing. Borrowing the idea from OFDM/LTE literature, where physical resource blocks are mapped to a time-frequency grid, an intuitive summary of those two beam probing classes and their visual abstraction can be seen in Fig. 5.

A. Iterative Beam Probing

After removing the cyclic prefix of all OFDM symbols, we can denote the input-output relation for subcarrier \( k \in 0, \ldots, K - 1 \) as

\[ y_k[n] = h_k f_{p,n} s_k[n] + z_k[n], n < N_{\text{coh}} \quad (9) \]

over one block of \( N_{\text{coh}} \) OFDM symbols. Here, \( z_k[n] \) is a circular complex Gaussian noise random variable with \( z_k[n] \sim \mathcal{CN}(0, 1) \) and \( s_k[n] \) denotes the transmit data symbols at subcarrier \( k \) and time instant \( n \) under OFDM symbol power constraint \( E_s \). \( f_{p,n} \) denotes one beamforming vector out of the codebook \( \mathcal{F}_p \) during OFDM symbol transmission \( n \in 1, \ldots, N_{\text{p}} \) (see Fig. 1 or Fig. 5a). The optimal beamformer is defined by the maximum mean achievable rate per subcarrier. Assuming CSI at the receiver and (9), the Rx needs to solve

\[ f_{p,n_{\text{max}}} = \arg \max_{f_{p,n} \in \mathcal{F}_p} \frac{1}{K} \sum_{k \in \mathcal{K}} \log_2 \left( 1 + \frac{E_s}{K \sigma_z^2} \| h_k f_{p,n} \|^2 \right), \quad (10) \]

As we cannot assume perfect channel knowledge of \( h_k \) during the initial beam alignment, the Rx relies on estimates to find the best beamformer from all time instants

\[ n_{\text{max}} = \arg \max_{n \in N_{\text{p}}} \sum_{k \in \mathcal{K}} y_k[n] y_k^*[n] \quad (11) \]

and deduces the corresponding \( f_{p,n_{\text{max}}} \) from the codebook \( \mathcal{F}_p \) due to the previously described time instant beamformer mapping. As we can only vary the beamformer from symbol to symbol, exhaustive search takes at least \( N_{\text{p}} = |\mathcal{F}_p| \) OFDM symbols to test all beamformers from the codebook \( \mathcal{F}_p \) to find the optimal one.

B. Parallel Beam Probing

Replacing \( f_{p,n} \) in (9) with the frequency dependent beamforming vector \( f_{p,k} \), we end up with the frequency-selective beam probing shown in Fig. 5b (or Fig. 2), where the input-output relation follows

\[ y_k[n] = h_k f_{p,k} s_k[n] + z_k[n], n < N_{\text{coh}}. \quad (12) \]
whereas in the iterative approach we need to wait at least the amount of beamformers to be tested. For example, we assume in Fig. 6 an OFDM scenario with $B = 2$GHz, $f_c = 60$GHz, $K = 2048$, $f_s = B/K$, $M_{Tx} = 8$, $d = c/(2f_c)$. OFDM symbol duration $T_s = 1/f_s$, and cluster size $L = 5$. All timing delays of these $L$ paths are drawn from a uniform distribution $\tau_f^{ch} \sim U(0, \tau_{cp})$, where $\tau_{cp} = 0.07T_s$ is the length of the cyclic prefix. Finally, we define the path with the shortest timing delay as the dominant path and by subtracting this timing offset $\tau_f^{ch} = \min_{l \in \{1, \ldots, L\}} \tau_f^{ch}$ from all other delays we can assume synchronization on the dominant path $l = 1$. The complex gain of path $l$ is $\alpha_l \sim C N(0, \sigma^{2}_{\alpha_l})$ and follows a circular complex Gaussian random variable with variance $\sigma^{2}_{\alpha_l}$. Hereby, $\sigma^{2}_{\alpha_l}$ is defined according to an exponential function such that a path with maximum delay $\tau_f^{ch}$ is attenuated by $-25$dB. Additionally, the channel is assumed to be passive meaning that $\sigma^{2}_{\alpha_l}$ is normalized such that $\sum_{l=1}^{L} \sigma^{2}_{\alpha_l} = 1$.

Now, we assume for the iterative approach the complete bandwidth $B$ or all subcarriers $k = 0, \ldots, K - 1$, respectively, are active for $N_p = 8$ OFDM symbols, whereas for the parallel test only $8$ subcarriers are active in one OFDM symbol $N_p = 1$. In this case the effective signal-to-noise ratio per subcarrier for the parallel test is higher than the one for the iterative test under equal OFDM symbol power constraint $E_s$. In Fig. 6, we plot the mean achievable rate $R$ from (15) for different $N_{coh} \in \{10, 20, 30\}$ with known $\mathbf{h}_k$, i.e., channel state information (CSI), and the mean achievable rate for estimated beamformers $\mathbf{f}_{p, k_{max}}$ and $\hat{\mathbf{f}}_{p, k_{max}}$, i.e., without channel state information (w/o. CSI). We can derive that the simulated results without CSI overlap with the results in case CSI is available. Furthermore, the parallel frequency search (Freq. S.) outperforms the iterative exhaustive search (Exh. S.) in terms of achievable data transmission rate due to high temporal efficiency of the beam alignment phase. This gain in performance is governed by the factor $N_p/N_{coh}$ and, therefore, applies to scenarios with large codebook sizes and short coherence times.

By simplifying the optimization problem from (10) to (13), we consequently lose performance in terms of maximum achievable rate $R$ in (15). This loss is due to the coherence bandwidth of the underlying mmWave channel, as we probe with the frequency-selective beam probing scheme each beamformer only on a very small fraction of the total bandwidth $B$. Therefore, the estimates get easily disturbed due to the frequency-selectivity of the channel and we decide for the wrong beamformer. We will address this issue using the same simulation parameters as previously described but increase the cluster size $L \in \{5, 10, 15\}$. Thus, the frequency-selectivity of the channel is increased and the risk of erroneous decisions rises. In Fig. 7, we plot the achievable rates normalized to the optimal mean achievable rate $R_{opt} = R_{exh. S. w. CSI}$ over the SNR for different $L$s. We can derive that for high SNR all the schemes (green and red curves) converge towards

\[
\begin{align*}
\hat{k}_{\text{max}} &= \arg\max_{k \in \mathcal{K}} \frac{E_s}{|k| \sigma^2_k} \| \mathbf{h}_k \mathbf{f}_{p, k} \|^2 & (13) \\
\hat{k}_{\text{max}} &= \arg\max_{k \in \mathcal{K}} \sum_{n=1}^{N_p} y_k[n] y_k^*[n]. & (14)
\end{align*}
\]

Assuming CSI at the receiver and using (12), the Rx is able to find the subcarrier

\[
\tilde{k}_{\text{max}} = \arg\max_{k \in \mathcal{K}} \frac{E_s}{|k| \sigma^2_k} \| \mathbf{h}_k \mathbf{f}_{p, k} \|^2
\]

with the largest instantaneous power. Considering that there is a demapping from $\tilde{k}_{\text{max}} \rightarrow \mathbf{f}_{p, \tilde{k}_{\text{max}}}$, the Rx can deduce its favourable beamformer $\mathbf{f}_{p, \tilde{k}_{\text{max}}}$ and informs the Tx. For unknown $\mathbf{h}_k$, the Rx estimates the optimal beamformer based on the magnitude of all channel outputs $y_k[n]$

\[
\hat{k}_{\text{max}} = \arg\max_{k \in \mathcal{K}} \frac{E_s}{|k| \sigma^2_k} \| \mathbf{h}_k \|^2
\]

To sum up, for iterative beam probing the mapping varies over time and for parallel beam probing the resource allocation changes over frequency (see Fig. 5). Therefore, one can easily derive the dualism between spending time or frequency resources to collect a sufficient number of observations. However, in the parallel approach estimates for all beamformers can be evaluated from the very beginning, whereas in the iterative approach we need to wait at least the amount of beamformers to be tested.

V. PERFORMANCE EVALUATION

For evaluation, we will compare the mean achievable rate per subcarrier

\[
R = \left(1 - \frac{N_p}{N_{coh}}\right) \frac{1}{K} \sum_{k \in \mathcal{K}} \log_2\left(1 + \frac{E_s}{K \sigma^2_k} \| \mathbf{h}_k \mathbf{f}_{p, k} \|^2\right)
\]

using the full bandwidth $B$ or all subcarriers $K$. Furthermore, we limit our discussion to beamformers from the orthogonal codebook $\mathbf{f}_p \in \mathcal{F}^+_p$. Beamformers found by the iterative approach (by solving (10) or demapping (11)) can directly be used in (15) and for beamformers found by the frequency-selective beam probing (by solving (13) or demapping (14)), we derive its corresponding wideband beamformer to evaluated (15).
their simulated achievable rates with CSI. Moreover, the performance penalty based on the channel realization gets also visible from Fig. 7, as the ratio \( R/R_{opt} \) for the blue plots (Freq. S. w. CSI) is not able to achieve 1. This backoff in maximum achievable rate is due to the limited number of used pilot subcarriers \([K]\) = 8, which leads to only a coarse sampling of the underlying channel transfer function \( h_k \). Nevertheless, considering that for mmWave channels we can assume \( L \) to be rather small or there is one dominant path, our proposal proves a robust beam alignment procedure and the performance degradation compared to optimal exhaustive search is negligible. As long as the frequency-selectivity of the underlying channel model is negligible, i.e., the channel response over the bandwidth \( B \) is constant and flat, parallel beam probing with few pilot subcarriers might be the preferred method for the initial beam alignment phase in future mmWave systems.

VI. Conclusion

In this work, we present a solution for the beam acquisition phase in mmWave communication systems using a novel frequency-selective beam probing network. The proposal is based on spectral analysis at the Rx knowing that certain frequencies correspond to beamformers from a given codebook \( \mathcal{F}_p \). Sharing such an instantaneous power profile with the transmitter, enables the Tx to assign beams such that the reliability of the link and the maximum achievable rate are increased. Moreover, the beam acquisition problem in the frequency domain can be solved in parallel for several beamformers. Therefore, such a parallel beam probing in frequency outperforms conventional beam acquisition schemes like exhaustive search in terms of latency. This comes at the cost of increased hardware complexity, which can be justifiable for rapid decisions in fast varying environments.

For future work, adding such pilot subcarriers constantly during the data transmission opens the possibility of closely tracking channel variations. Moreover, frequency-selective beam probing offers the potential for tracking users in different directions at the same time. With close tracking channel variations and scheduling the users accordingly, the maximum achievable sum-rate in multi-user scenarios can potentially be increased in fast-time varying environments.

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