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Soumendu Ghosh

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SPC Lab, Department of ECE
Indian Institute of Science

- **Introduction**

1. Virtual full-duplex cooperative non-orthogonal multiple access (NOMA) framework for a downlink two-hop network assisted by multiple half-duplex decode and forward (DF) relay stations (RSs).
2. VFD: a set of RS receives the signal from the BS and at the same time transmits the received signal of the previous instant to the mobile station.
3. They have proposed a RS selection algorithm with adaptive inter-RS interference management.
4. Simulation results show that the proposed RS selection algorithm outperforms the existing algorithms in terms of both the outage probability and the DMT, which has the best performance reported in the literature.

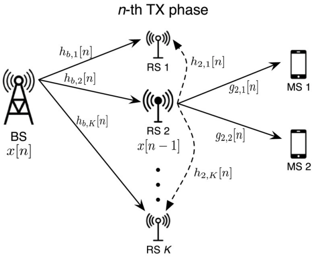

Figure 1: System Model (A single BS, two MSs, and K half-duplex RSs in the absence of the direct link from the BS to two MSs.).

- Assume a virtual full-duplex operation at the RSs, in which a particular RS sends a packet while the other RSs receive a packet from the BS at the same time.
- The number of total successive transmission phases is assumed to be N.
- The received signal at the k -th RS in the n -th transmission phase is given by

$$y_k^r[n] = h_{b,k}[n]x[n] + h_{j,k}[n]x[n-1] + z_k^r[n], 1 \leq k \leq K, 1 \leq n \leq N$$

- In NOMA, the BS sends the superimposed signal that is given by $x[n] = \sqrt{a_1}s_1[n] + \sqrt{a_2}s_2[n]$
- At the i -th MS, the received signal is given by

$$y_i^m[n] = g_{j,i}[n]x[n-1] + z_i^m[n],$$

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$$\text{(C1)}: \ \log\left(1+\frac{a_1\left|h_{b,k}[n]\right|^2}{a_2\left|h_{b,k}[n]\right|^2+1/\rho}\right) \geq \frac{NR_1}{N-1},$$

$$\text{(C2)}: \ \log\left(1+a_2\rho|h_{b,k}[n]|^2\right) \geq \frac{NR_2}{N-1},$$

- **Theorem:** Assuming K half-duplex RSs between the BS and two MSs, the outage probability of the VFD cooperative NOMA technique is given by

$$\Pr\{\mathcal{O}\} = \sum_{t=0}^{K}\left(1-\exp\left(-\frac{2^{\frac{NR_2}{N-1}}-1}{a_2\rho}\right)\right)^t \pi_t,$$

where $\pi_t, \forall t \in \{0,\ldots,K\}$ denotes a $(K+1)$-dimensional stationary distribution vector for the Markov chain. Thus,$\boldsymbol{\pi} \triangleq [\pi_0, \pi_1, \ldots, \pi_K]$ and $\sum_{i=0}^{K}\pi_i = 1$.

Spatial Covariance Estimation for Millimeter Wave Hybrid Systems Using Out-of-Band Information: Anum Ali, Nuria González-Prelcic, Robert W. Heath, Jr.

- **Introduction:**

1. In high mobility applications of mmWave communications frequent link configuration can be a source of significant overhead.
2. Sub-6 GHz channel covariance as an out-of-band side information for mmWave link configuration.
3. They have proposed an out-of-band covariance translation approach and an out-of-band aided compressed covariance estimation approach.
   The out-of-band covariance translation eliminates the in-band training completely, whereas out-of-band aided covariance estimation relies on in-band as well as out-of-band training.

- **System Model**

1. OFDM transmission with K sub-carriers.
   The transmission symbols on sub-carrier k are denoted as $\mathsf{s}[k] \in \mathbb{C}^{N_\mathrm{s} \times 1}$, and follow $\mathbb{E}[\mathsf{s}[k]\mathsf{s}^*[k]] = \frac{P}{KN_\mathrm{s}}\mathbf{I}_{N_\mathrm{s}}$.

- If $H[k]$ denotes the frequency-domain $N_{RX} \times N_{TX}$ mmWave MIMO channel on sub-carrier k , then the post-processing received signal on sub-carrier k can be represented as

$$\begin{aligned} \mathsf{y}[k] =& \mathbf{W}^*_{\mathrm{BB}}[k]\mathbf{W}^*_{\mathrm{RF}}\mathsf{H}[k]\mathbf{F}_{\mathrm{RF}}\mathbf{F}_{\mathrm{BB}}[k]\mathsf{s}[k] + \mathbf{W}^*_{\mathrm{BB}}[k]\mathbf{W}^*_{\mathrm{RF}}\mathsf{n}[k], \\ =& \mathsf{W}^*[k]\mathsf{H}[k]\mathsf{F}[k]\mathsf{s}[k] + \mathsf{W}^*[k]\mathsf{n}[k], \end{aligned}$$

- MIMO channel matrix $H[d]$ can be written as

$$\mathbf{H}[d] = \sqrt{N_{\mathrm{RX}} N_{\mathrm{TX}}} \sum_{c=1}^{C} \sum_{r_c=1}^{R_c} \alpha_{r_c} p(dT_{\mathrm{s}} - \tau_c - \tau_{r_c}) \times \mathbf{a}_{\mathrm{RX}}(\theta_c + \vartheta_{r_c}) \mathbf{a}^*_{\mathrm{TX}}(\phi_c + \varphi_{r_c}),$$

  The MIMO channel at sub-carrier k ,$H[k]$ can be expressed as

$$\mathsf{H}[k] = \sum_{d=0}^{D-1} \mathbf{H}[d] e^{-\mathrm{j}\frac{2\pi k}{K}d},$$

- The transmit covariance of the channel on sub-carrier k is defined as $\mathsf{R}_{\mathrm{TX}}[k] = \frac{1}{N_{\mathrm{RX}}}\mathbb{E}[\mathsf{H}^*[k]\mathsf{H}[k]]$ while the receive covariance is $\mathsf{R}_{\mathrm{RX}}[k] = \frac{1}{N_{\mathrm{TX}}}\mathbb{E}[\mathsf{H}[k]\mathsf{H}^*[k]]$.
- Estimating $\mathsf{R} \in \mathbb{C}^{N_{\mathrm{RX}} \times N_{\mathrm{RX}}}$ from $\underline{\mathsf{R}} \in \mathbb{C}^{\underline{N}_{\mathrm{RX}} \times \underline{N}_{\mathrm{RX}}}$.
  We assume that the estimate of the sub-6 GHz covariance $\hat{\underline{\mathsf{R}}}$ is available.

- Under the small AS assumption, the channel covariance can be written as

$$[\mathsf{R}]_{i,j} = e^{\mathrm{j}(\mathrm{i}-\mathrm{j})2\pi\Delta\sin(\theta)}\Phi\big((i-j)2\pi\Delta\cos(\theta)\sigma_{\vartheta}\big).$$

- Now, under the assumption of uncorrelated clusters, the total covariance can be written as

$$\underline{\mathsf{R}} = \sum_{\underline{c}=1}^{\underline{C}} \underline{\epsilon}_{\underline{c}}\underline{\mathsf{R}}(\underline{\theta}_{\underline{c}}, \underline{\sigma}_{\underline{\vartheta},\underline{c}}) + \underline{\sigma}_{\underline{\mathsf{n}}}^2\mathbf{I}.$$

- The mmWave covariance corresponding to the cth cluster is denoted as $\mathsf{R}(\underline{\theta}_{\underline{c}}, \underline{\sigma}_{\underline{\vartheta},\underline{c}})$. Similar to sub-6 GHz covariance $\underline{\mathsf{R}}(\underline{\theta}_{\underline{c}}, \underline{\sigma}_{\underline{\vartheta},\underline{c}})$ the mmWave covariance $\mathsf{R}(\underline{\theta}_{\underline{c}}, \underline{\sigma}_{\underline{\vartheta},\underline{c}})$ is also calculated using the expressions in Table I.

TABLE I

THEORETICAL EXPRESSIONS FOR COVARIANCE $[\mathsf{R}]_{i,j}$

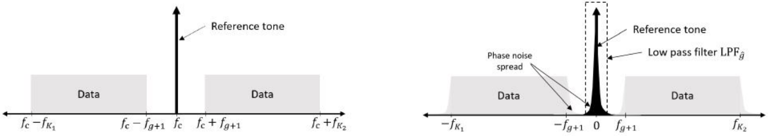
PAS	Expression
Truncated Laplacian [36]	$\frac{\beta e^{\mathrm{j}2\pi\Delta(i-j)\sin(\theta)}}{1+\frac{\sigma_{\vartheta}^2}{2}[2\pi\Delta(i-j)\cos(\theta)]^2}, \ \beta = \frac{1}{1-e^{-\sqrt{2}\pi/\sigma_{\vartheta}}}$
Truncated Gaussian [35]	$e^{-((i-j)2\pi\Delta\cos(\theta)\sigma_{\vartheta})^2} e^{\mathrm{j}2\pi\Delta(i-j)\sin(\theta)}$
Uniform [35]	$\frac{\sin((i-j)\varrho_{\vartheta})}{((i-j)\varrho_{\vartheta})} e^{\mathrm{j}2\pi\Delta(i-j)\sin(\theta)}, \ \varrho_{\vartheta} = \sqrt{3}\times 2\pi\Delta\sigma_{\vartheta}\cos(\theta)$

- **Introduction**

1. A special case of hybrid-beamforming: Analog beamforming [*only one up/down conversion chain.*]
2. It is most power efficient and lowest cost implementation.
3. **Major Drawback:** Since one down-conversion chain has to be time multiplexed across the Rx antennas for CE, several pilot re-transmissions are required ($o(M_{Tx}M_{Rx})$).
4. This overhead increases the system latency and makes the initial access process cumbersome.
5. Using only analog hardware and avoiding time multiplexing reduces the overhead: analog channel estimation (ACE). However PLL aided recovery limits the performance due to high Rx phase noise.
6. Generalized ACE:
    - **CACE:** avoid PLL, compensate oscillator phase noise and exploit both amplitude and the phase information of the channel.
    - **PACE:** prevents wastage of the transmit resources on a continuous reference, however, uses PLL.
    - **MA-FSR:** resilient to phase noise, low hardware cost, but poor bandwidth efficiency.

- **CACE enabled Rx.**

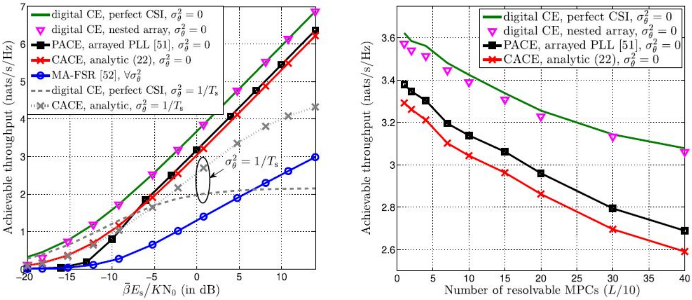
1. In CACE, a reference tone, i.e. a sinusoidal tone at a known frequency, is continuously transmitted along with the data by the TX.
2. At the RX, the received signal at each antenna is converted to base-band by a bank of mixers and a local oscillator that is tuned (approximately) to the reference frequency.
3. These filtered outputs, which are implicit estimates of the channel response (including amplitude and phase) at the reference frequency, are then used as control signals to a variable gain, analog phase-shifter array to generate the RX analog beam.
4. The un-filtered base-band received signals at each antenna are processed by these phase shifters, added and fed to a single ADC for demodulation.



(a) Transmit signal at TX antenna m (b) Baseband signal at RX antenna m

Figure 2: Illustration in OFDM receiver

- **Key Results:**



(a) Sparse channel

(b) Dense stochastic channel

Figure 3: Throughput of ACE schemes (PACE, CACE, MA-FSR) and of digital CE versus SNR and L

1. As is evident from Fig. a, PACE and CACE suffer only a $\leq$ 2dB beamforming loss in compared to digital CE in sparse channels and above a threshold SNR.
2. While CACE performs marginally worse than PACE at high SNR due to power wastage on a continuous reference, unlike PACE it does not suffer from PLL based carrier recovery losses at low SNR.
3. As observed in Fig. b, the performance of ACE schemes degrades slightly faster with L than of digital CE.

1. Censored Spectrum Sharing Strategy for MIMO Systems in Cognitive Radio Networks: Jyoti Mansukhani and Priyadip Ray.
2. Asymmetric Modulation Design for Wireless Information and Power Transfer With Nonlinear Energy Harvesting: Ekaterina Bayguzina and Bruno Clerckx.
3. Effective Secrecy Rate for a Downlink NOMA Network: Wenjuan Yu et al.
4. Capacity Scaling in a Non-Coherent Wideband Massive SIMO Block Fading Channel: Felipe Gómez-Cuba et al.
5. Fairness and Sum-Rate Maximization via Joint Subcarrier and Power Allocation in Uplink SCMA Transmission: Joao V. C. Evangelista et al.

Thank You