Spatial Modulation for MIMO Wireless Systems

Marco Di Renzo(1), Harald Haas(2) and Ali Ghrayeb(3)

(1) Laboratory of Signals and Systems (L2S), CNRS – SUPÉLEC – University of Paris-Sud XI
3 rue Joliot-Curie, 91192 Gif-sur-Yvette, France
marco.direnzo@lss.supelec.fr

(2) The University of Edinburgh, Institute for Digital Communications (IDCOM)
Mayfield Road, Edinburgh, EH9 3JL, UK
h.haas@ed.ac.uk

(3) Concordia University, Department of Electrical and Computer Engineering
1455 de Maisonneuve West, Montreal, H3G 1M8, Canada
aghrayeb@ece.concordia.ca

http://www.fp7-greenet.eu

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Spatial Modulation for Generalized MIMO: Challenges, Opportunities, and Implementation

This tutorial paper offers a comprehensive overview of the state of the art in spatial modulation for generalized multiple-input–multiple-output (MIMO) technologies.

By Marco Di Renzo, Member IEEE, Harald Haas, Member IEEE, Ali Ghayeb, Senior Member IEEE, Shinya Sugiura, Senior Member IEEE, and Lajos Hanzo, Fellow IEEE

This paper has supplementary downloadable material available at http://ieeexplore.ieee.org, provided by the authors. This includes three multimedia MP4 360p format movie clips, which show a description of the working principle of spatial modulation (SMintro.mp4), some experimental activities related to spatial modulation (SMtestbed.mp4), and a tutorial presentation that summarizes the main content of the present paper (SMtutorial.mp4). These files SMintro.mp4, SMtestbed.mp4, and SMtutorial.mp4 are 17.70, 13.40, and 334.00 MB in size, respectively. Furthermore, two presentations in PDF format are available, which provide a 95-slide short (SM_ShortPresentation.pdf) and a 421-slide comprehensive (SM_LongPresentation.pdf) description of spatial modulation and of the research challenges described in this paper. These files SM_ShortPresentation.pdf and SM_LongPresentation.pdf are 3.50 and 16.40 MB in size, respectively.

YouTube:

- Spatial Modulation
  (http://www.youtube.com/watch?v=cPKIbxrEDho)
- The Advantages of Spatial Modulation
  (http://www.youtube.com/watch?v=baKkBxzf4fY)
- The World's First Spatial Modulation Demonstration
  (http://www.youtube.com/watch?v=a6yUuJFUtZ4)

- Tutorial
  (http://www.youtube.com/watch?v=cNgCJK4oimM&feature=youtu.be)
Spatial Modulation for Generalized MIMO: Challenges, Opportunities and Implementation
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Why MIMO?

- Array gain (beamforming), spatial division multiple access
- Spatial multiplexing: Rate $= \min(N_t, N_r) \log_2(1+\text{SNR})$
- Reliability: BEP $\sim \text{SNR}^{-(N_tN_r)}$
Regardless of the use as diversity or spatial multiplexing system, the main drawback of conventional MIMO systems is the increased complexity, increased power/energy consumption, and high cost. Why?

- **Inter-channel interference (ICI):** Introduced by coupling multiple symbols in time and space – signal processing complexity.
- **Inter-antenna synchronization (IAS):** Detection algorithms require that all symbols are transmitted at the same time.
- **Multiple radio frequency (RF) chains:** RF elements are expensive, bulky, no simple to implement, and do not follow Moore’s law.
- **Energy consumption:** The energy efficiency decreases linearly with the number of active antennas (RF chains) and it mostly depends on the Power Amplifiers (>60%) – EARTH model.
Conventional vs. Single-RF MIMO

The Energy Efficiency (EE) Challenge (1/3)


MIMO Gain WITHOUT Considering Circuit Power

Fig. 6. MIMO vs. SISO EE gain against the SE when considering the theoretical PCM.

Fig. 8. MIMO vs. SISO EE gain against the number of antenna elements $n_{ant} = t = r$ when considering the theoretical PCM.

MIMO Gain Considering Circuit Power

Fig. 7. MIMO vs. SISO EE gain against the SE for different types of BS when considering the double linear PCM of (6).

Fig. 9. MIMO vs. SISO EE gain against the number of antennas $n_{int} = t = r$ for different types of BS when considering the double linear PCM of (6).

SE vs. EE Tradeoff (1/2)

- SE-oriented systems are designed to maximize the capacity under peak or average power constraints, which may lead to transmitting with the maximum allowed power for long periods, thus deviate from EE design.

- EE is commonly defined as information bits per unit of transmit energy. It has been studied from the information-theoretic perspective for various scenarios.

- For an additive white Gaussian noise (AWGN) channel, it is well known that for a given transmit power, $P$, and system bandwidth, $B$, the channel capacity is:

$$R = \left(\frac{1}{2}\right) \log_2 \left(1 + \frac{P}{N_0 B}\right) = \left(\frac{1}{2}\right) \eta_{SE}$$

- bits per real dimension or degrees of freedom (DOF), where $N_0$ is the noise power spectral density. According to the Nyquist sampling theory, DOF per second is $2B$. Therefore, the channel capacity is $C = 2BR$ b/s. Consequently, the EE is:

$$\eta_{EE} = \frac{C}{P} = \frac{2R}{N_0 \left(2^{2R} - 1\right)} = \frac{\eta_{SE}}{N_0 \left(2^{\eta_{SE}} - 1\right)}$$

- It follows that the EE decreases monotonically with $R$ (i.e., with SE).

SE vs. EE Tradeoff (2/2)

Now, Imagine a New Modulation for MIMOs:

- Having one (or few) active RF chains but still being able to exploit all transmit-antenna elements for multiplexing and transmit-diversity gains

- Offering Maximum-Likelihood (ML) optimum decoding performance with single-stream decoding complexity

- Working without the need of (power inefficient) linear modulation schemes (QAM) or allowing us to use constant-envelope modulation (PSK) with negligible performance degradation

Spatial Modulation (SM) has the inherent potential to meet these goals
**SM – In a Nutshell**

- **Spatial Multiplexing**
  - Vertical Bell Laboratories Layered Space-Time

- **Orthogonal Space-Time-Block Coding**
  - Transmit Diversity
  - Orthogonal Space-Time-Block Coding

- **Spatial Modulation**
  - Spatial Modulation
  - S2 = 0/1
SM – How It Works (3D Constellation Diagram)

SM – How It Works (2/3)

... 1110 0001 ...

Spatial Constellation

SM – How It Works (3/3)

... 1110 0001 ...

SM – Transmitter

Binary Source

… 100101011100110010 101001010100111010111 …

SM Mapper

\[ \log_2(N_t) + \log_2(M) \]

Antenna Selection

Tx2

Signal Selection

-1 (BPSK)
SM – Wireless Channel

Communication Channel

Wireless Channel

Tx0  Tx1  Tx2  Tx3

Rx
SM - Receiver

Rx a priori CSI

Detection

D0(+) = distance(Rx,+Tx0)
D0(-) = distance(Rx,-Tx0)
D1(+) = distance(Rx,+Tx1)
D1(-) = distance(Rx,-Tx1)
D2(+) = distance(Rx,+Tx2)
D2(-) = distance(Rx,-Tx2)
D3(+) = distance(Rx,+Tx3)
D3(-) = distance(Rx,-Tx3)

Compute min\{Di(±)\}
Common Misunderstandings

- What is the difference with Transmit Antenna Selection (TAS)?
  - TAS is closed-loop (transmit-diversity). SM is open-loop (spatial-multiplexing).
  - In TAS, antenna switching depends on the end-to-end performance. In SM, antenna switching depends on the incoming bit-stream.

- SIMO: $\log_2(M)$ bpcu – MIMO: $N_t\log_2(M)$ bpcu – SM: $\log_2(N_t)+\log_2(M)$ bpcu. So, SM is spectral efficiency (SE) sub-optimal. Why using it?
  - Correct. But what about signal processing complexity, cost, total power consumption, and energy efficiency (EE)?
  - Are we looking for SE-MIMO? For EE-MIMO? Or for a good SE/EE tradeoff?

- SM needs many more transmit-antennas than conventional MIMO for the same SE. Is the comparison fair? Is having so many antennas practical?
  - What does fair mean? Same transmit-antennas? Same RF chains?
  - What about massive MIMO? What about mm-Wave communications?

- Due to the encoding mechanism, is SM more sensitive to channel estimation errors than conventional MIMO?
  - No, it is as/more robust as/than MIMO and we have results proving it.
Our Proposal: Single-RF Large-Scale SM-MIMO

- The rationale behind SM–MIMO communications for the design of spectral and energy efficient cellular networks is based upon two main pillars:

1) **Minimize**, given some performance constraints, the number of active antenna–elements in order to increase the EE by reducing the circuit power consumption (single–RF MIMO principle).

2) **Maximize**, given some implementation and size constraints, the number of passive antenna–elements in order to increase both the SE and the EE by reducing the transmit power consumption (large–scale MIMO principle). This is realized by capitalizing on the multiplexing gain introduced by the “spatial-constellation diagram”.

G. Wright “GreenTouch Initiative: Large Scale Antenna Systems Demonstration”, 2011 Spring meeting, Seoul, South Korea. Available at: http://www.youtube.com/watch?v=U3euDDr0uvo.

Massive MIMO (2/5)

- With very large MIMO, we think of systems that use antenna arrays with an order of magnitude more elements than in systems being built today, say a hundred antennas or more.

- Very large MIMO entails an unprecedented number of antennas simultaneously serving a much smaller number of terminals.

- In very large MIMO systems, each antenna unit uses extremely low power, of the order of mW.

- As a bonus, several expensive and bulky items, such as large coaxial cables, can be eliminated altogether. (The coaxial cables used for tower-mounted base stations today are up to four centimeters in diameter).

- Very-large MIMO designs can be made extremely robust in that the failure of one or a few of the antenna units would not appreciably affect the system. Malfunctioning individual antennas may be hotswapped.


Massive MIMO (3/5)

- The main effect of scaling up the dimensions is that uncorrelated thermal noise and fast fading can be averaged out and vanish so that the system is predominantly limited by interference from other transmitters.

- If we could assign an orthogonal pilot sequence to every terminal in every cell then large numbers of base station antennas would eventually defeat all noise and fading, and eliminate both intra-and inter-cell interference.

- But there are not enough orthogonal pilot sequences for all terminals. Pilot sequences have to be reused. The performance of a very large array becomes limited by interference arising from re-using pilots in neighboring cells (pilot contamination problem).

- With an infinite number of antennas, the simplest forms of user detection and precoding, i.e., matched filtering (MF) and eigenbeamforming, become optimal.

- Spectral efficiency is independent of bandwidth, and the required transmitted energy per bit vanishes.

Consider a MIMO Multiple Access (MAC - UPLINK) system with $N$ antennas per BS and $K$ users per cell:

$$y = \sum_{k=1}^{K} h_k x_k + n$$

where channel and noise are i.i.d. RVs with zero mean and unit variance.

By the strong law of large numbers:

$$\frac{1}{N} h_m^H y \xrightarrow{N \to \infty \text{ and } K = \text{const}} x_m$$

Thus, with an unlimited number of BS antennas:

- Uncorrelated interference and noise vanish
- The matched filter is optimal
- The transmit power can be made arbitrarily small

Massive MIMO (5/5) – In Formulas

- Assume now that transmitter $m$ and $j$ use the same pilot:

$$\hat{h}_m = h_m + h_j + n_m$$

- Thus, by the strong law of large numbers:

$$\frac{1}{N} \hat{h}_m^H y \xrightarrow{N \to \infty \text{ and } K=\text{const}} x_m + x_j$$

- Thus, with an unlimited number of BS antennas:
  - Uncorrelated interference, noise, and estimation errors vanish
  - The performance of the matched filter receiver is limited by pilot contamination
  - Matched filter and minimum mean square receivers provide the same limiting performance

Massive MIMO vs. SM-MIMO (downlink)

- **Massive MIMO:**
  - Many (hundreds or more) transmit-antennas
  - All transmit-antennas are simultaneously-active: multi-RF MIMO but antennas are less expensive and more EE than state-of-the-art
  - EE: reduction of transmit (RF) power

- **SM-MIMO:**
  - Many (hundreds or more) transmit-antennas
  - One (or few) transmit-antennas are simultaneously-active: single-RF MIMO
  - EE: reduction of transmit (RF) power and circuits power
Power-Amplifier Aware MISO Design

**Motivation:**
... the mutual information was maximized under an assumption of a **limited output power**. However, in many applications it is desirable to instead limit the **total consumed power**, consisting of both output power and losses in the transmitter **chain**. The scientific literature agrees that the power amplifier is the largest source of losses in the transmitter.

**Contribution:**
... we utilize the analytic expression of amplifier losses to design MIMO beamforming schemes. We observe that our MISO solution, differently from the traditional MRT beamforming, is such that **some antennas in general are turned off**.

**Takeaway Message:**
The proposed procedure allows us to **turn off antennas while operating optimally**, which is **beneficial in cases where dissipated power per antenna is significant**. This also gives us the possibility to turn off whole radio frequency chains with filters and mixers, which saves additional power.

Transmission Concepts Related to SM (1/5)

New multiple antenna designs based on compact parasitic architectures have been proposed to enable multiplexing gains with a single active RF element and many passive antenna elements. The key idea is to change the radiation pattern of the array at each symbol time instance, and to encode independent information streams onto angular variations of the far-field in the wave-vector domain.

New MIMO schemes jointly combining multiple-antenna transmission and Automatic Repeat reQuest (ARQ) feedback have been proposed to avoid keeping all available antennas on, thus enabling MIMO gains with a single RF chain and a single power amplifier. This solution is named Incremental MIMO. The main idea is to reduce complexity and to improve the energy efficiency by having one active antenna at a time, but to exploit ARQ feedback to randomly cycle through the available antennas at the transmitter in case of incorrect data reception.

New directional modulation schemes for mm-Wave frequencies have been proposed to enable secure and low-complexity wireless communications. The solution is named Antenna Subset Modulation (ASM). The main idea in ASM is to modulate the radiation pattern at the symbol rate by driving only a subset of antennas in the array. While randomly switching antenna subsets does not affect the symbol modulation for a desired receiver along the main direction, it effectively randomizes the amplitude and phase of the received symbol for an eavesdropper along a sidelobe.

In Millimeter–wave Mobile Broadband (MMB) system design, the cost of implementing one RF chain per transmit–antenna can be prohibitive. For this reason, analog baseband beamforming or RF beamforming with one or a few active RF chains can be promising low–complexity solutions. Proposal: low–complexity hybrid RF/baseband precoding schemes where large antenna–arrays are driven by a limited number of transmit/receive chains.

To Summarize: SM-MIMO Advantages…

- Higher throughput
- Simpler receiver design
- Simpler transmitter design
- Lower transmit power supply
- Better efficiency of the power amplifiers

… and Some Disadvantages/Trade-Offs

- Spectral efficiency sub-optimality
- Fast antenna switching
- Time-limited pulse shaping
- Favorable propagation conditions
- Training overhead
- Directional beamforming (for mmWave applications)

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A Glimpse into the History of SM


Research Groups Working on SM

- University of Edinburgh, UK (H. Haas)
- CNRS – SUPELEC – University of Paris-Sud XI, France (M. Di Renzo)
- Concordia University, Canada (A. Ghrayeb)
- University of Tabuk, Saudi Arabia (R. Y. Mesleh)
- University of Southampton, UK (L. Hanzo)
- Princeton University, US (V. Poor)
- Istanbul Technical University, Turkey (E. Basar, E. Panayirci)
- Tokyo University, Japan (S. Sugiura)
- Indian Institute of Science, India (K. V. S. Hari and A. Chockalingam)
- Québec University - INRS, Canada (S. Aissa)
- The University of Akron, US (H. R. Bahrami)
- Academia Sinica, Taiwan (a large group)
- Tsinghua University and many other universities, China (many groups)
- Le Quy Don Technical University, Vietnam (T. X. Nam)
- etc., etc., etc…

Collaborations with: Univ. of L’Aquila (Italy), CTTC (Spain), Univ. of Bristol (UK), Heriot-Watt Univ. (UK), EADS (Germany), etc…
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Spatial Modulation (SM)

Space Shift Keying (SSK)

- Information is conveyed only by the Spatial-Constellation diagram
  - No signal modulation $\rightarrow$ more efficient power amplifiers (no linearity constraints)
  - Simplified demodulation
  - Larger antenna-arrays are needed for the same spectral efficiency

\[
b = \begin{bmatrix} b_1 & b_2 \end{bmatrix} \quad \text{symbol} \quad \text{antenna index } j \quad x = \begin{bmatrix} x_1 & \cdots & x_4 \end{bmatrix}^T
\]

\[
\begin{bmatrix} 0 & 0 \\ 0 & 1 \\ 1 & 0 \\ 1 & 1 \end{bmatrix} \quad \begin{bmatrix} 0 \\ 1 \\ 2 \\ 3 \end{bmatrix} \quad \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}^T
\]

\[
y = Hx_l + n = h^l + n
\]

Transmitter Design – Encoding (3/8)

Generalized SM and SSK

- SM and SSK are appealing because of their single RF design which greatly simplifies the transmitter.

- However, their rates are:
  - \( \log_2(N_t) + \log_2(M) \) bpcu for SM
  - \( \log_2(N_t) \) bpcu for SSK

- Rate and complexity can be traded-off by allowing more than one active antenna in each time instance, as well as by allowing different numbers of active antennas per time slots:
  - Generalized SSK
  - Generalized SM
  - Variable Generalized SSK/SM
### Generalized SSK (GSSK)

<table>
<thead>
<tr>
<th>( b = [b_1 \ b_2 \ b_3] )</th>
<th>( j )</th>
<th>( x = [x_1 \ x_2 \ \ldots \ x_5]^T )</th>
</tr>
</thead>
<tbody>
<tr>
<td>([0 \ 0 \ 0])</td>
<td>(1,2)</td>
<td>( \begin{bmatrix} \frac{1}{\sqrt{2}} &amp; \frac{1}{\sqrt{2}} &amp; 0 &amp; 0 &amp; 0 \end{bmatrix}^T )</td>
</tr>
<tr>
<td>([0 \ 0 \ 1])</td>
<td>(1,3)</td>
<td>( \begin{bmatrix} \frac{1}{\sqrt{2}} &amp; 0 &amp; \frac{1}{\sqrt{2}} &amp; 0 &amp; 0 \end{bmatrix}^T )</td>
</tr>
<tr>
<td>([0 \ 1 \ 0])</td>
<td>(1,4)</td>
<td>( \begin{bmatrix} \frac{1}{\sqrt{2}} &amp; 0 &amp; 0 &amp; \frac{1}{\sqrt{2}} &amp; 0 \end{bmatrix}^T )</td>
</tr>
<tr>
<td>([0 \ 1 \ 1])</td>
<td>(1,5)</td>
<td>( \begin{bmatrix} \frac{1}{\sqrt{2}} &amp; 0 &amp; 0 &amp; 0 &amp; \frac{1}{\sqrt{2}} \end{bmatrix}^T )</td>
</tr>
<tr>
<td>([1 \ 0 \ 0])</td>
<td>(2,3)</td>
<td>( \begin{bmatrix} 0 &amp; \frac{1}{\sqrt{2}} &amp; \frac{1}{\sqrt{2}} &amp; 0 &amp; 0 \end{bmatrix}^T )</td>
</tr>
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<td>([1 \ 0 \ 1])</td>
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</tr>
<tr>
<td>([1 \ 1 \ 0])</td>
<td>(2,5)</td>
<td>( \begin{bmatrix} 0 &amp; \frac{1}{\sqrt{2}} &amp; 0 &amp; 0 &amp; \frac{1}{\sqrt{2}} \end{bmatrix}^T )</td>
</tr>
<tr>
<td>([1 \ 1 \ 1])</td>
<td>(3,4)</td>
<td>( \begin{bmatrix} 0 &amp; 0 &amp; \frac{1}{\sqrt{2}} &amp; \frac{1}{\sqrt{2}} &amp; 0 \end{bmatrix}^T )</td>
</tr>
</tbody>
</table>

**Rate** = \( \log_2 \left( \frac{N_t}{n_t} \right) \)

**Rate** = 3 bpcu

\( N_t = 5 \)

\( n_t = 2 \)

### Generalized SM (GSM)

<table>
<thead>
<tr>
<th>Grouped Bits</th>
<th>Antenna Combination ((\ell))</th>
<th>Symbol ((s))</th>
</tr>
</thead>
<tbody>
<tr>
<td>0000</td>
<td>(1,2)</td>
<td>-1</td>
</tr>
<tr>
<td>0001</td>
<td>(1,2)</td>
<td>+1</td>
</tr>
<tr>
<td>0010</td>
<td>(1,3)</td>
<td>-1</td>
</tr>
<tr>
<td>0011</td>
<td>(1,3)</td>
<td>+1</td>
</tr>
<tr>
<td>0100</td>
<td>(1,4)</td>
<td>-1</td>
</tr>
<tr>
<td>0101</td>
<td>(1,4)</td>
<td>+1</td>
</tr>
<tr>
<td>0110</td>
<td>(1,5)</td>
<td>-1</td>
</tr>
<tr>
<td>0111</td>
<td>(1,5)</td>
<td>+1</td>
</tr>
<tr>
<td>1000</td>
<td>(2,3)</td>
<td>-1</td>
</tr>
<tr>
<td>1001</td>
<td>(2,3)</td>
<td>+1</td>
</tr>
<tr>
<td>1010</td>
<td>(2,4)</td>
<td>-1</td>
</tr>
<tr>
<td>1011</td>
<td>(2,4)</td>
<td>+1</td>
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<td>-1</td>
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<td>1101</td>
<td>(3,5)</td>
<td>+1</td>
</tr>
<tr>
<td>1110</td>
<td>(4,5)</td>
<td>-1</td>
</tr>
<tr>
<td>1111</td>
<td>(4,5)</td>
<td>+1</td>
</tr>
</tbody>
</table>

Rate = \(4\text{bpcu}\)

\(N_t = 5\)

\(n_t = 2\)

BPSK

\[
\text{Rate} = \left\lfloor \log_2 \left( \frac{N_t}{n_t} \right) \right\rfloor + \log_2 (M)
\]

Transmitter Design – Encoding (6/8)

Variable Generalized SSK/SM (VGSM/VGSSK)

<table>
<thead>
<tr>
<th>Grouped Bits</th>
<th>Antenna Combination ($\gamma^{\text{VGSM}}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>000</td>
<td>(1)</td>
</tr>
<tr>
<td>001</td>
<td>(2)</td>
</tr>
<tr>
<td>010</td>
<td>(3)</td>
</tr>
<tr>
<td>011</td>
<td>(4)</td>
</tr>
<tr>
<td>100</td>
<td>(1,2)</td>
</tr>
<tr>
<td>101</td>
<td>(1,3)</td>
</tr>
<tr>
<td>110</td>
<td>(1,4)</td>
</tr>
<tr>
<td>111</td>
<td>(2,3)</td>
</tr>
</tbody>
</table>

Rate = 3bpcu + $\log_2(M)$

$N_t = 4$

$n_t = 1$ and $2$

MQAM/MPSK

\[
\begin{align*}
\text{Rate}_{\text{VGSM}} &= \log_2(M) + \left\lfloor \log_2 \left( \sum_{n_t=1}^{N_t} \left( \begin{array}{c} N_t \\ n_t \end{array} \right) \right) \right\rfloor \\
&= \log_2(M) + \left\lfloor 2^{N_t} - 1 \right\rfloor = \log_2(M) + (N_t - 1) \\
\text{Rate}_{\text{VGSSK}} &= \left\lfloor \log_2 \left( \sum_{n_t=0}^{N_t} \left( \begin{array}{c} N_t \\ n_t \end{array} \right) \right) \right\rfloor = \left\lfloor 2^{N_t} \right\rfloor = N_t
\end{align*}
\]
Amalgamating SM and Spatial Multiplexing?

Transmitter Design – Encoding (7/8)
Transmitter Design – Encoding (8/8)

Reasoning on the Tradeoffs

- **Performance**
  \[ \text{PEP} \propto Q\left( \sqrt{\text{SNR} \left\| \mathbf{H}_{x_k} - \mathbf{H}_{x_h} \right\|^2} \right) \]

- **Signal processing complexity (detection)**

- **Total vs. active (RF chains) number of transmit-antennas**

\[ R = \frac{\text{Complexity GSM}}{\text{Complexity SM}} \]
Outline

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The first proposed demodulator for SM is based on a two-step approach:

- Detection of the antenna index (spatial-constellation diagram)
- Detection of the modulated symbol (signal-constellation diagram)

\[
\hat{l} = \arg \max_l \left\{ \frac{h_l^H y}{\| h_l \|_F^2} \right\}
\]

\[
\hat{s} = \arg \min_s \left\{ \| y - h_{\hat{l}}^H s \|_F^2 \right\}
\]

Maximum-Likelihood (ML) optimum decoding:

- Spatial- and signal-constellation diagrams are jointly decoded

\[
\left( \hat{l}, \hat{s} \right) = \arg\min_{l,s} \left\{ \| y - H x_{l,s} \|_F^2 \right\}
\]

\[
= \arg\min_{l,s} \left\{ \| y - h_l x_s \|_F^2 \right\}
\]

Many other sub-optimal demodulators have been proposed recently.

In general, they offer a trade-off between complexity and performance.

Sometimes, they provide good performance for low/medium SNRs, while they performance degrades for high SNRs.

We consider two examples:

- The application of Compressed Sensing to SM
- The application of Sphere Decoding to SM


Compressed Sensing (CS) Generalized Space Shift Keying

- The idea is to leverage the inherent sparsity of SSK modulation: the number of active antennas is much less than the radiating elements ($n_t < N_t$)

- SSK demodulation is re-formulated as a convex program via CS

- CS-SSK uses 1-norm metric instead of 2-norm of ML demodulation

- The demodulation complexity is:

  \[
  \begin{cases}
    \text{ML} : O\left( N_r N_t^{n_t} \right) \\
    \text{CS} : O\left( N_r N_t n_t \right)
  \end{cases}
  \]

Compressed Sensing (CS) Generalized Space Shift Keying

\[
\begin{cases}
y = \sqrt{\rho} Hx + n \\
y \in \mathbb{C}^{N_r \times 1}, \ x \in \mathbb{R}^{N_t \times 1}, \ H \in \mathbb{C}^{N_r \times N_t}, \ n \in \mathbb{C}^{N_r \times 1}
\end{cases}
\]

- The idea is to leverage the inherent sparsity of SSK modulation: the number of active antennas is much less than the radiating elements \((n_t < N_t)\).
- \(x\) can be re-constructed with high probability by 1-norm minimization, as follows:

\[
\hat{x} = \arg \min_{y=\Phi x} \left\{ \| x \|_1 \right\}
\]

Compressed Sensing (CS) Generalized Space Shift Keying

where:

- \( \Phi \) is an \( N_r \times N_t \) that satisfies the Restricted Isometric Property (RIP). CS theory says that, with high probability, matrix \( \Phi \) can be obtained by generating its elements from a Normal distribution with zero mean and variance \( 1/N_r \). The RIP ensures that pairs of columns of \( \Phi \) are orthogonal to each other with high probability.

- The number of observations \( N_r \) should be chosen as follows:
  \[
  N_r = O \left( n_t \log_2 \left( \frac{N_t}{n_t} \right) \right)
  \]

- The authors use Orthogonal Matching Pursuit (OMP). The idea is find the non-zero elements of \( x \) by computing the correlation \( \Phi^T y \). If \( \Phi \) satisfies the RIP, then \( \Phi^T \Phi \) is nearly orthonormal and the largest coefficients of \( \Phi^T y \) correspond to the non-zero coefficients of \( x \).

Sphere Decoding (SD) Spatial Modulation

- Optimum detector based on the ML principle:

\[
[\ell_t^{\text{(ML)}}, s_t^{\text{(ML)}}] = \arg \min_{\ell \in \{1, 2, \ldots, N_t\}, s \in \{s_1, s_2, \ldots, s_M\}} \left\{ \| y - h_{\ell} s \|_F^2 \right\}
\]

\[
= \arg \min_{\ell \in \{1, 2, \ldots, N_t\}, s \in \{s_1, s_2, \ldots, s_M\}} \left\{ \sum_{r=1}^{N_r} |y_r - h_{\ell,r}s|^2 \right\}
\]

- Computational complexity of ML (real multiplications):

\[
C_{\text{ML}} = 8N_r N_t M
\]

since evaluating each Euclidean distance requires 8 real multiplications

The SD algorithm avoids an exhaustive search by examining only those points that lie inside a sphere of radius $R$:

$$
\gamma(N_r, 2\alpha N_r) = \Gamma(N_r)(1 - \varepsilon)
$$

$$
R = 2\alpha N_r \sigma_N^2
$$

Three sphere decoders for SM are proposed and studied:

1. **Rx-SD**, which aims at reducing the receive search space

\[
\left[ \ell_t^{(\text{Rx-SD})}, S_t^{(\text{Rx-SD})} \right] = \arg \min_{\ell \in \{1, 2, \ldots, N_t\}, \quad \{s \in \{s_1, s_2, \ldots, s_M\} \quad N_r(\ell, s) = N_r} \{ \sum_{r=1}^{N_r(\ell, s)} |y_r - h_{\ell, r} s|^2 \}
\]

2. **Tx-SD**, which aims at reducing the transmit search space

\[
\left[ \ell_t^{(\text{Tx-SD})}, S_t^{(\text{Tx-SD})} \right] = \arg \min_{(\ell, s) \in \Theta_R} \left\{ \sum_{r=1}^{N_r} |y_r - h_{\ell, r} s|^2 \right\}
\]

3. **C-SD**, which aims at reducing both transmit and receive search spaces

\[
\left[ \ell_t^{(\text{C-SD})}, S_t^{(\text{C-SD})} \right] = \arg \min_{(\ell, s) \in \Theta_R \quad N_r(\ell, s) = N_r} \left\{ \sum_{r=1}^{N_r} |y_r - h_{\ell, r} s|^2 \right\}
\]

Rx–SD searches the paths leading to each point \((l,s)\) as long as it is still inside the sphere when adding up the signals at each receive-antenna.

Sphere Decoding (SD) Spatial Modulation – Tx-SD

\[
\begin{align*}
\left[ \ell_t^{(\text{Tx-SD})}, s_t^{(\text{Tx-SD})} \right] &= \arg \min_{(\ell, s) \in \Theta_R} \left\{ \sum_{r=1}^{N_r} | y_r - h_{\ell,r} s_r |^2 \right\} \\
\Theta_R &= \left\{ (\ell, s) \text{ with } \ell \in \{1, 2, \ldots N_t \} \text{ and } s \in \{ s_1, s_2, \ldots s_M \} | \| \tilde{y} - \tilde{H} \bar{x}_{\ell,s} \|_F^2 \leq R^2 \right\} \\
&= \left\{ (\ell, s) \text{ with } \ell \in \{1, 2, \ldots N_t \} \text{ and } s \in \{ s_1, s_2, \ldots s_M \} | \sum_{i=1}^{2N_t} \left( \bar{z}_i - \sum_{j=i}^{2N_t} \bar{p}_{i,j} \bar{x}_{j} (\ell, s) \right)^2 \leq R_Q^2 \right\}
\end{align*}
\]

\[\frac{-R_Q + \bar{z}_i}{\bar{p}_{i,i}} \leq \bar{x}_i (\ell, s) \leq \frac{R_Q + \bar{z}_i}{\bar{p}_{i,i}}\]

\[\frac{-R_Q + \bar{z}_{i,i+N_t}}{\bar{p}_{i,i}} \leq \bar{x}_i (\ell, s) \leq \frac{R_Q + \bar{z}_{i,i+N_t}}{\bar{p}_{i,i}}\]

The C–SD is a two–step detector that works as follows:

1. First, the Tx–SD algorithm is used to reduce the transmit search space. The subset of points $\Theta_R$ is computed.

2. Second, the Rx–SD algorithm is used to reduce the receive search space. More specifically, Rx–SD is applied only on the subset of points $\Theta_R$ computed in Step 1.

$$\begin{align*}
[\ell_t^{(C-SD)}, s_t^{(C-SD)}] &= \arg\min_{(\ell, s) \in \Theta_R} \left\{ \sum_{r=1}^{\tilde{N}_r} |y_r - h_{\ell,r}s|^2 \right\} 
\end{align*}$$

Sphere Decoding (SD) Spatial Modulation – C-SD

- The complexity of Rx–SD is:
  \[ C_{Rx-SD} = 8 \sum_{\ell=1}^{N_t} \sum_{s=1}^{M} \tilde{N}_r(\ell, s) \]

- The complexity of Tx–SD is:
  \[ C_{Tx-SD} = C_{\Theta_R} + 8N_r \text{card}\{\Theta_r\} \]

- The complexity of Cx–SD is:
  \[ C_{C-SD} = C_{\Theta_R} + 8 \sum_{(\ell, s) \in \Theta_R} \tilde{N}_r(\ell, s) \]

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6bpcu – i.i.d. Rayleigh fading

Error Performance – Numerical Results (2/24)

8bpcu – i.i.d. Rayleigh fading

3bpcu – i.i.d. Rayleigh fading – $N_r=4$

1bpcu and 3bpcu – i.i.d. Rayleigh fading – $N_r=2$

1bpcu and 3bpcu – i.i.d. Rayleigh fading – $N_r=1, 2, 4$

Error Performance – Numerical Results (9/24)

3bpcu – i.i.d. Rayleigh fading – $N_r=4$

Error Performance – Numerical Results (10/24)

8bpcu – i.i.d. Rayleigh fading – N_r=4

GSM: N_t = 12, n_t = 3
VGSM: N_t = 8, M = 2
SM: N_t = 128, M = 2
SMX: N_t = 8, M = 2

8bpcu – Rayleigh fading, exponential correlation ($\beta=0.6$) – $N_t=4$

GSM: $N_t = 12$, $n_t = 3$
VGSM: $N_t = 8$, $M = 2$
SM: $N_t = 128$, $M = 2$
SMX: $N_t = 8$, $M = 2$

8bpcu – i.i.d. Rician fading – $N_r=4$

GSM: $N_t = 12$, $n_t = 3$
VGSM: $N_t = 8$, $M = 2$
SM: $N_t = 128$, $M = 2$
SMX: $N_t = 8$, $M = 2$

8bpcu – Rician fading, exponential correlation ($\beta=0.6$) – $N_r=4$

GSM: \(N_t = 12, n_t = 3\)
VGSM: \(N_t = 8, M = 2\)
SM: \(N_t = 128, M = 2\)
SMX: \(N_t = 8, M = 2\)

Error Performance – Numerical Results (14/24)

i.i.d. Rayleigh fading

$N_t = 256$

$n_t = 1$

Varying $N_r$

i.i.d. Rayleigh fading

Setup “2”:
N_t = 256, n_t = 2, N_r = 16

Setup “3”:
N_t = 64, n_t = 3, N_r = 24

Error Performance – Numerical Results (16/24)

i.i.d. Rayleigh fading

Setup “CS2-16”:
N_t = 256, n_t = 2, N_r = 16

Setup “CS2-20”:
N_t = 256, n_t = 2, N_r = 20

Setup “CS3-24”:
N_t = 64, n_t = 3, N_r = 24

Setup “CS3-30”:
N_t = 64, n_t = 3, N_r = 30

Error Performance – Numerical Results (19/24)

i.i.d. Rayleigh fading – $N_t=4$, $N_r=4$

Error Performance – Numerical Results (22/24)

Single-RF vs. Multi-RF (SSK vs. Spatial-Multiplexing MIMO)

Error Performance – Numerical Results (23/24)

Single-RF vs. Multi-RF (SM vs. Spatial-Multiplexing MIMO)

Single-RF vs. Multi-RF (SM vs. Spatial-Multiplexing MIMO)

i.i.d. Rayleigh fading
8 bpcu
N_r=4

Figure 8.3  Average SEP of 16-QAM over a Nakagami-$m$ channel versus the average SNR per symbol.

Error Performance – Main Trends (2/38)

\[
\text{ABEP} \leq \text{ABEP}^{\text{CUB}} = \frac{2}{N_t \log_2 (N_t)} \sum_{i_1=1}^{N_t} \sum_{i_2=i_1+1}^{N_t} N(i_1, i_2) \text{APEP}(\text{TX}_i \to \text{TX}_{i_1})
\]

\[
\text{APEP}(\text{TX}_i \to \text{TX}_{i_1}) = \frac{1}{\pi} \int_0^{\pi/2} M_{\gamma_{i_1,2}} \left( \frac{E_u/4N_0}{2 \sin^2(\theta)} \right) d\theta
\]

\[
M_{\gamma_{i_1,2}}(s) = \sum_{k=0}^{+\infty} \frac{AC^{m+2k-1}}{2^{m-1}4^k (k!) \Gamma(k+m)} \Psi_k(s)
\]

\[
\Psi_k(s) = \frac{A}{4} (s + B_1)^{-(m+k)} (s + B_2)^{-(m+k)} G_{2,2}^{1,2} \left( \frac{s^2}{(s + B_1)(s + B_2)} \right) \left| \begin{array}{cc} 1-m-k & 1-m-k \\ 1 & 0 \\ 0 & 0 \end{array} \right|
\]

\[
A = \frac{4m^{m+1}}{\Gamma(m)\Omega_1\Omega_2(1-\rho_{12}^2\beta_1^2\beta_2^2)(\sqrt{\Omega_1\Omega_2\rho_{12}^2\beta_1^2\beta_2^2})^{m-1}}
\]

\[
\left\{B_i\right\}^2 = \frac{m}{\Omega_i(1-\rho_{12}^2\beta_1^2\beta_2^2)} ; C = \frac{2m\sqrt{\rho_{12}^2\beta_1^2\beta_2^2}}{\sqrt{\Omega_1\Omega_2(1-\rho_{12}^2\beta_1^2\beta_2^2)}}
\]

Fig. 1. Comparison between Monte Carlo simulation (markers) and analytical model (solid lines). $2 \times 1$ MISO system. Uncorrelated fading model in (16) (i.e., $\rho_{12}^2 = 0$) with balanced power (i.e., $\Omega_1 = \Omega_2 = 1$), and $m_2 = 2$. 
Fig. 3. Comparison between Monte Carlo simulation (markers) and analytical model (solid lines). 2 × 1 MISO system. Correlated fading model in (21) (i.e., $\rho \beta_1^2 \beta_2^2 = 0.25$ on the left and $\rho \beta_1^2 \beta_2^2 = 0.75$ on the right) with balanced power (i.e., $\Omega_1 = \Omega_2 = 1$).
Fig. 2. Comparison between Monte Carlo simulation (markers) and analytical model (solid lines). 2 × 1 MISO system. Uncorrelated fading model in (16) (i.e., $\rho_{\beta_1\beta_2} = 0$) with unbalanced power (i.e., $\Omega_1 = 10\Omega_2$), and $m_2 = 2$, $\Omega_2 = 1$. 

Error Performance – Main Trends (5/38)
Fig. 4. Comparison between Monte Carlo simulation (markers) and analytical model (solid lines). 2 × 1 MISO system. Correlated fading model in (21) (i.e., $\rho_{\beta_1^2\beta_2^2} = 0.25$) with unbalanced power (i.e., $\Omega_1 = 10\Omega_2$), and $\Omega_2 = 1$. 

Error Performance – Main Trends (6/38)
Fig. 5. Comparison between Monte Carlo simulation (markers) and analytical model (solid lines). $2 \times 1$ MISO system. Correlated fading model in (21) (i.e., $\rho_{\beta_1^2 \beta_2^2} = 0.75$) with unbalanced power (i.e., $\Omega_1 = 10 \Omega_2$), and $\Omega_2 = 1$. 
Fig. 6. Comparison between Monte Carlo simulation (markers) and analytical model (solid lines). $2 \times 1$ MISO system. Correlated fading model in (28).

Scenario a: $m_1 = m_2 = 1$, $\Omega_1 = \Omega_2 = 2/3$, $\delta_1 = 0.50$, $\delta_2 = \delta_3 = \delta_4 = 0.45$. Scenario b: $m_1 = 1$, $m_2 = 2.5$, $\Omega_1 = 1$, $\Omega_2 = 10$, $\delta_1 = \delta_2 = \delta_3 = 0.45$, $\delta_4 = -0.45$. Scenario c: $m_1 = m_2 = 1$, $\Omega_1 = 2/3$, $\Omega_2 = 20/3$, $\delta_1 = 0.50$, $\delta_2 = \delta_3 = \delta_4 = 0.45$. 

Error Performance – Main Trends (8/38)
Fig. 11. Comparison between Monte Carlo simulation and analytical model. 8 × 1 MISO system. Correlated fading model in (21) (i.e., \( \{ \rho_{\beta_i \beta_j}^2 \}_{i,j=1}^8 \approx \{ \rho_{\beta_i \beta_j} \}_{i,j=1}^8 = \exp(-0.5|i-j|) \)) with balanced (i.e., \( \{ \Omega_i \}_{i=1}^8 = 1 \)) and unbalanced (i.e., \( \Omega_1 = 1, \{ \Omega_i \}_{i=2}^8 = 3(i-1) \)) power, and \( \{ m_i \}_{i=1}^8 = 2 \).
Error Performance – Main Trends (10/38)

Fig. 9. Performance comparison (analytical model only) between 4 × 1 and 8 × 1 MISO systems for various fading scenarios. Uncorrelated fading model in (16) (i.e., \( \rho_{ij}^2 = 0 \)). Scenario a: \( N_t = 4, \{m_i\}_i^N_t = i, \{\Omega_i\}_i^N_t = 1 \). Scenario b: \( N_t = 4, \{m_i\}_i^N_t = i, \Omega_1 = 1, \{\Omega_i\}_i^N_t = 4(i - 1) \). Scenario c: \( N_t = 4, \{m_i\}_i^N_t = 5 - i, \Omega_1 = 1, \{\Omega_i\}_i^N_t = 4(i - 1) \). Scenario d: \( N_t = 8, \{m_i\}_i^N_t = i, \{\Omega_i\}_i^N_t = 1 \). Scenario e: \( N_t = 8, \{m_i\}_i^N_t = i, \Omega_1 = 1, \{\Omega_i\}_i^N_t = 3(i - 1) \). Scenario f: \( N_t = 8, \{m_i\}_i^N_t = 9 - i, \Omega_1 = 1, \{\Omega_i\}_i^N_t = 3(i - 1) \).
Error Performance – Main Trends (11/38)

\[
\text{PEP} (t_1 \rightarrow t_2) = \frac{2^{N_r-1} \prod_{r=1}^{N_r} \frac{m_{t_1,r} \cdot m_{t_2,r}}{m_{t_1,r} \cdot m_{t_2,r}} \left( \frac{m_{t_1,r}}{m_{t_2,r}} \right)^{(m_{t_1,r} + m_{t_2,r} - 1)} \sqrt{\frac{\Gamma(N_r + 1)}{(N_r + 1)}} \Gamma(N_r + 1/2) \Gamma(m_{t_1,r} + m_{t_2,r} - 1) \left( \frac{E_m}{4N_0} \right)^{-N_r}}{\sqrt{\pi}}
\]

\[
\text{PEP} (t_1 \rightarrow t_2) = \frac{2^{N_r-1} \left( \sum_{k_1=0}^{\infty} \frac{C_{t_1,t_2,r}^{m_r + 2k_r - 1} k_{N_r}^{k_r}}{2^{m_r - 1 + k_r} k_{N_r}^{k_r} \Gamma(k_{N_r} + m_r)} \right) \Gamma(N_r + 1/2) \Gamma(N_r + 1/2) \left( \frac{E_m}{4N_0} \right)^{-N_r}}{\sqrt{\pi}}
\]

\[N_t = 8\]
\[m = 2.5\]
\[\Omega = 1\]

Figure 8.8 Combined effect of Nakagami-$n$ (Rice) fading and imperfect synchronization on the average bit error probability of BPSK.

\[ \text{ABEP}_{2,1} = \int_{0}^{+\infty} Q\left(\sqrt{\gamma}\zeta\right) f_{\gamma_{2,1}}(\zeta) \, d\zeta \]

\[ = \frac{\Psi_{2,1}}{2\sqrt{\pi}} \sum_{k=0}^{+\infty} \frac{\vartheta_{k}^{(2,1)}}{\Gamma(\nu(2,1) + k)} G_{2,2}^{2,1}\left(\frac{\gamma_{1}(2,1)}{2} \left| \begin{array}{ccc} 1 - \nu(2,1) - k & 1 \\ 0 & 1/2 \end{array} \right. \right) \]

\[ \Psi_{2,1} = \prod_{h=1}^{L} \left[ \frac{\chi_{h}^{(2,1)}}{\chi_{h}^{(2,1)}} \exp\left(-\frac{\kappa_{h}^{(2,1)}}{\chi_{h}^{(2,1)}}\right) \right] \]

\[ \vartheta_{k}^{(2,1)} = \frac{1}{k} \sum_{r=1}^{k} \left\{ \sum_{b=1}^{L} \nu_{r}^{(2,1)} \left(1 - \frac{\chi_{1}^{(2,1)}}{\chi_{b}^{(2,1)}}\right)^{r} + \frac{r\chi_{1}^{(2,1)} \kappa_{r}^{(2,1)}}{\chi_{b}^{(2,1)}^{2}} \left(1 - \frac{\chi_{1}^{(2,1)}}{\chi_{b}^{(2,1)}}\right)^{r-1}\right\} \vartheta_{k-r}^{(2,1)} \]

\[ \text{ABEP}_{2,1} = \frac{1}{\pi} \int_{0}^{\pi/2} M_{\gamma_{2,1}}\left(\frac{\gamma}{2\sin^{2}(\theta)}\right) \, d\theta \]

\[ M_{\gamma_{2,1}}(s) = \prod_{h=1}^{L} \left[ \left(1 + \chi_{h}^{(2,1)} s\right)^{-\nu_{h}^{(2,1)}} \exp\left(-\frac{\kappa_{h}^{(2,1)}}{1 + \chi_{h}^{(2,1)} s}\right) \right] \]

Fig. 1. SSK modulation: ABEP against $E_m/N_0$. Solid, dashed, and dotted lines denote the analytical model in Section III and markers Monte Carlo simulations. Setup: i) $N_t = 2$, ii) $N_r = 2$, iii) $\Omega_{i,l} = 10$dB and $K_{R}^{(i,l)} = K_R$ for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$, and iv) $\rho = 0.00$ (solid lines), $\rho = 0.25$ (dashed lines), $\rho = 0.75$ (dotted lines).
Fig. 3. SSK modulation: ABEP against $E_m/N_0$. Solid, dashed, and dotted lines denote the analytical model in Section III and markers Monte Carlo simulations. Setup: i) $N_t = 4$, ii) $N_r = 2$, iii) $\Omega_{i,l} = 10$dB and $K_R^{(i,l)} = K_R$ for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$, and iv) $\rho = 0.00$ (solid lines), $\rho = 0.25$ (dashed lines), $\rho = 0.75$ (dotted lines).
Fig. 4. SSK modulation: ABEP against $E_m/N_0$. Solid, dashed, and dotted lines denote the analytical model in Section III and markers Monte Carlo simulations. Setup: i) $N_t = 4$, ii) $N_r = 4$, iii) $\Omega_{i,l} = 10$dB and $K^{(i,l)}_R = K_R$ for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$, and iv) $\rho = 0.00$ (solid lines), $\rho = 0.25$ (dashed lines), $\rho = 0.75$ (dotted lines).
Proposition 1: The ABEP in (3) can be tightly upper bounded as follows:

\[ \text{ABEP} \leq \text{ABEP}_{\text{signal}} + \text{ABEP}_{\text{spatial}} + \text{ABEP}_{\text{joint}} \]

where \( \text{ABEP}_{\text{signal}} \), \( \text{ABEP}_{\text{spatial}} \), and \( \text{ABEP}_{\text{joint}} \) are defined as

\[
\begin{align*}
\text{ABEP}_{\text{signal}} &= \frac{1}{N_t} \log_2(M) \sum_{n_t=1}^{N_t} \text{ABEP}_{\text{MOD}}(n_t) \\
\text{ABEP}_{\text{spatial}} &= \frac{1}{M} \log_2(N_t) \sum_{l=1}^{M} \text{ABEP}_{\text{SSK}}(l) \\
\text{ABEP}_{\text{joint}} &= \frac{1}{N_t M} \log_2(N_t M) \sum_{n_t=1}^{N_t} \sum_{l=1}^{M} \sum_{\tilde{n}_t \neq n_t} \sum_{\tilde{l} \neq l} \\
&\quad \times \{ N_H(\tilde{n}_t \rightarrow n_t) + N_H(\chi_{\tilde{l}} \rightarrow \chi_l) \} \quad \forall (n_t, \chi_l, \tilde{n}_t, \chi_{\tilde{l}}) 
\end{align*}
\]

\[ \text{ABEP}_{\text{MOD}}(n_t) = \frac{1}{M} \frac{1}{\log_2(M)} \sum_{l=1}^{M} \sum_{\tilde{l}=1}^{M} \times \left[ N_H(\chi_{\tilde{l}} \rightarrow \chi_l) E_{\alpha(n_t)} \left\{ \Pr \left\{ \chi_{\tilde{l}} = \chi_l \mid \chi_{\tilde{l}} \right\} \right\} \right] \]
\[ \text{ABEP}_{\text{SSK}}(l) = \frac{1}{N_t} \frac{1}{\log_2(N_t)} \sum_{n_t=1}^{N_t} \sum_{\tilde{n}_t=1}^{N_t} \times \left[ N_H(\tilde{n}_t \rightarrow n_t) \Psi_l(n_t, \tilde{n}_t) \right] \]

\[ \Psi_l(n_t, \tilde{n}_t) = \left( \frac{1}{\pi} \right) \int_{0}^{\pi/2} M_{\gamma(n_t, \tilde{n}_t)} \left( \tilde{\gamma} \kappa_l^2 / 2 \sin^2(\theta) \right) d\theta ; \]
\[ \Upsilon(n_t, \chi_l, \tilde{n}_t, \chi_{\tilde{l}}) = \left( \frac{1}{\pi} \right) \int_{0}^{\pi/2} M_{\gamma(n_t, \chi_l, \tilde{n}_t, \chi_{\tilde{l}})} \left( \tilde{\gamma} / 2 \times \sin^2(\theta) \right) d\theta . \]

The diversity order over Nakagami-m fading channels is:

$$\text{Div}_{SM} = \min \left\{ N_r, m_{Nak} N_r \right\}$$

- If $m_{Nak} > 1$, $\text{Div}_{SM} = N_r$, the ABEP is dominated by the spatial-constellation diagram
- If $m_{Nak} < 1$, $\text{Div}_{SM} = m_{Nak} N_r$, the ABEP is dominated by the signal-constellation diagram
- $\text{Div}_{SIMO} = m_{Nak} N_r$ for every $m_{Nak}$

The diversity order over Rician fading channels is:

$$\text{Div}_{SM} = N_r$$

Error Performance – Main Trends (21/38)

i.i.d. Rayleigh Fading

\[
\begin{align*}
\text{ABEP}_{\text{signal}} &= 2^{\text{log}_2(M)} \text{ABEP}_{\text{Rayleigh}} \\
\text{ABEP}_{\text{spatial}} &= \frac{1}{N_t} \sum_{l=1}^{M} \mathcal{R} \left( 4\sigma_0^2 \tilde{\gamma} \kappa_l^2 \right) \\
\text{ABEP}_{\text{joint}} &= \frac{1}{N_t} \sum_{l=1}^{M} \sum_{l' \neq l=1}^{M} \left[ \left( \frac{N_t \log_2(N_t)}{2} + (N_t - 1) N_H (\chi_l \rightarrow \chi_l) \right) \mathcal{R} \left( 2\sigma_0^2 \tilde{\gamma} \left( \kappa_l^2 + \kappa_{l'}^2 \right) \right) \right]
\end{align*}
\]

\[
\mathcal{R} (\xi) = \left[ \frac{1}{2} \left( 1 - \sqrt{\frac{\xi}{2+\xi}} \right) \right]^{N_r} \sum_{n_r=0}^{N_r-1} \left( \begin{array}{c}
N_r - 1 - r \\
r
\end{array} \right) \left[ \frac{1}{2} \left( 1 + \sqrt{\frac{\xi}{2+\xi}} \right) \right]^{n_r}
\]

High-SNR

\[
\begin{align*}
\text{ABEP}_{\text{spatial}} &\geq \frac{N_t}{2} \frac{1}{M} \frac{\log_2(N_t)}{\log_2(N_t M)} 2^{-N_r} \left( \frac{2N_r-1}{N_r} \right) \Theta^{(M,N_r)}_{\text{spatial}} \left( 4\sigma_0^2 \tilde{\gamma} \right)^{-N_r} \\
\text{ABEP}_{\text{joint}} &\geq \frac{N_t}{2} \frac{1}{M} \frac{\log_2(N_t)}{\log_2(N_t M)} \left( \frac{2N_r-1}{N_r} \right) \Theta^{(M,N_r)}_{\text{joint}} + \frac{N_t-1}{M} \frac{1}{\log_2(N_t M)} \left( \frac{2N_r-1}{N_r} \right) \Theta^{(M,N_r,H)}_{\text{joint}} \left( 4\sigma_0^2 \tilde{\gamma} \right)^{-N_r}
\end{align*}
\]

Error Performance – Main Trends (22/38)

i.i.d. Rayleigh Fading

\[ \Delta_{\text{SNR}}^{(X/Y)} : \text{SNR gain of Y compared to X} \]

\[
\Delta_{\text{SNR}}^{(X/Y)} = 10 \log_{10} \left( \frac{\text{SNR}_X}{\text{SNR}_Y} \right) = -\left( \frac{10}{N_r} \right) \log_{10} \left( \Pi_{\text{SNR}}^{(X/Y)} \right)
\]

<table>
<thead>
<tr>
<th>Formula</th>
<th>Description</th>
</tr>
</thead>
</table>
| \( \Pi_{\text{SNR}}^{(\text{PSK}/\text{SM-PSK})} \) | \[
\Pi_{\text{SNR}}^{(\text{PSK}/\text{SM-PSK})} = \left\{ \frac{N_t M}{2} \log_2(N_t) + 2^{-N_r} \log_2(M) \right\} G_{\text{PSK}}^{\text{MOD}} \left( \frac{M^{\text{PSK}}}{z} \right) - N_r G_{\text{PSK}}^{\text{MOD}} \left( \frac{M^{\text{PSK}}}{z} \right)
\]
| \( \Pi_{\text{SNR}}^{(\text{QAM}/\text{SM-QAM})} \) | \[
\Pi_{\text{SNR}}^{(\text{QAM}/\text{SM-QAM})} = \frac{1}{M} \left[ \frac{N_t}{2} \log_2(N_t) \Theta_{\text{spatial}}^{(M,N_r)} + \frac{N_t}{2} \log_2(N_t) \Theta_{\text{joint}}^{(M,N_r)} \right] + (N_t - 1) 2^{N_r} \Theta_{\text{joint}}^{(M,N_r)} + \log_2(M) \left[ G_{\text{QAM}}^{\text{MOD}}(I_M) + G_{\text{QAM}}^{\text{MOD}}(J_M) \right]
\]
| \( \Pi_{\text{SNR}}^{(\text{SSK}/\text{SM-QAM})} \) | \[
\Pi_{\text{SNR}}^{(\text{SSK}/\text{SM-QAM})} = \frac{1}{M} \left[ \frac{N_t}{2} \log_2(N_t) \Theta_{\text{spatial}}^{(M,N_r)} + \frac{N_t}{2} \log_2(N_t) \Theta_{\text{joint}}^{(M,N_r)} \right] + (N_t - 1) 2^{N_r} \Theta_{\text{joint}}^{(M,N_r)} + \log_2(M) \left[ G_{\text{QAM}}^{\text{MOD}}(I_M) + G_{\text{QAM}}^{\text{MOD}}(J_M) \right]
\]
| \( \Pi_{\text{SNR}}^{(\text{SSK}/\text{QAM})} \) | \[
\Pi_{\text{SNR}}^{(\text{SSK}/\text{QAM})} = \frac{G_{\text{QAM}}^{\text{MOD}}(I_M) + G_{\text{QAM}}^{\text{MOD}}(J_M)}{2^{N_{\text{SSK}}} \log_2(N_{\text{SSK}})}
\]
| \( \Pi_{\text{SNR}}^{(\text{SSK}/\text{PSK})} \) | \[
\Pi_{\text{SNR}}^{(\text{SSK}/\text{PSK})} = \frac{2^{-N_r} G_{\text{PSK}}^{\text{MOD}}(M^{\text{PSK}})}{2^{N_{\text{SSK}}} \log_2(N_{\text{SSK}})}
\]

### Error Performance – Main Trends (23/38)

#### i.i.d. Rayleigh Fading

<table>
<thead>
<tr>
<th>$N_r = 1$</th>
<th>2 bpcu</th>
<th>3 bpcu</th>
<th>4 bpcu</th>
<th>5 bpcu</th>
<th>6 bpcu</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rate $(R) / \Delta^{(X/Y)}_{SNR}$</td>
<td>2 bpcu</td>
<td>3 bpcu</td>
<td>4 bpcu</td>
<td>5 bpcu</td>
<td>6 bpcu</td>
</tr>
<tr>
<td>(PSK, SM–PSK)</td>
<td>−2.4304</td>
<td>−0.9691</td>
<td>0.5799</td>
<td>2.0412</td>
<td>3.2906</td>
</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>−1.5761</td>
<td>0.2803</td>
<td>2.2640</td>
<td>4.2597</td>
</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>N.A.</td>
<td>0.0684</td>
<td>2.1512</td>
<td>4.3511</td>
</tr>
<tr>
<td>(QAM, SM–QAM)</td>
<td>−2.4304</td>
<td>−1.0939</td>
<td>−3.3199</td>
<td>−2.2055</td>
<td>−4.1422</td>
</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>−1.7009</td>
<td>−2.7542</td>
<td>−3.3406</td>
<td>−4.3064</td>
</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>N.A.</td>
<td>−2.9661</td>
<td>−2.3416</td>
<td>−4.9156</td>
</tr>
<tr>
<td>(SSK, SM–QAM)</td>
<td>0.5799</td>
<td>0.7918</td>
<td>−0.2854</td>
<td>0.2460</td>
<td>−0.3481</td>
</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>0.1848</td>
<td>0.2803</td>
<td>−0.8890</td>
<td>−0.5123</td>
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<tr>
<td></td>
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<td>N.A.</td>
<td>0.0684</td>
<td>0.1100</td>
<td>−1.1215</td>
</tr>
</tbody>
</table>

### Error Performance – Main Trends (24/38)

i.i.d. Rayleigh Fading

<table>
<thead>
<tr>
<th>Rate / $\Delta^{(X/Y)}_{\text{SNR}}$</th>
<th>2 bpcu</th>
<th>3 bpcu</th>
<th>4 bpcu</th>
<th>5 bpcu</th>
<th>6 bpcu</th>
</tr>
</thead>
<tbody>
<tr>
<td>(PSK, SM–PSK)</td>
<td>−1.0543</td>
<td>1.9011</td>
<td>4.5154</td>
<td>5.6931</td>
<td>5.9642</td>
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<tr>
<td>N.A.</td>
<td>1.6453</td>
<td>5.3471</td>
<td>8.8845</td>
<td>11.1650</td>
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</tr>
<tr>
<td>N.A.</td>
<td>N.A.</td>
<td>5.2585</td>
<td>9.2429</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(QAM, SM–QAM)</td>
<td>−1.0543</td>
<td>1.7709</td>
<td>0.1040</td>
<td>2.3751</td>
<td>0.9242</td>
</tr>
<tr>
<td>N.A.</td>
<td>1.5152</td>
<td>2.0064</td>
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<tr>
<td>N.A.</td>
<td>N.A.</td>
<td>1.9177</td>
<td>4.2581</td>
<td></td>
<td>2.5484</td>
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<tr>
<td>(SSK, SM–QAM)</td>
<td>0.4509</td>
<td>0.3959</td>
<td>−1.7622</td>
<td>−1.8280</td>
<td>−3.6242</td>
</tr>
<tr>
<td>N.A.</td>
<td>0.1401</td>
<td>0.1401</td>
<td>−1.9196</td>
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</tr>
<tr>
<td>N.A.</td>
<td>N.A.</td>
<td>0.0515</td>
<td>0.0550</td>
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<td>−2.0000</td>
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</table>

### Error Performance – Main Trends (25/38)

**i.i.d. Rayleigh Fading**

<table>
<thead>
<tr>
<th>Rate / $\Delta^{(X/Y)}_{\text{SNR}}$</th>
<th>$N_r = 3$</th>
<th>2 bpcu</th>
<th>3 bpcu</th>
<th>4 bpcu</th>
<th>5 bpcu</th>
<th>6 bpcu</th>
</tr>
</thead>
<tbody>
<tr>
<td>(PSK, SM–PSK)</td>
<td>$-0.6461$</td>
<td>3.0103</td>
<td>5.5248</td>
<td>5.9627</td>
<td>6.0094</td>
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</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>2.8560</td>
<td>7.2677</td>
<td>10.9352</td>
<td>11.9378</td>
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</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>N.A.</td>
<td>7.2144</td>
<td>11.8624</td>
<td>16.1295</td>
<td></td>
</tr>
<tr>
<td>(QAM, SM–QAM)</td>
<td>$-0.6461$</td>
<td>2.7651</td>
<td>0.9978</td>
<td>3.3520</td>
<td>1.6807</td>
<td></td>
</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>2.6108</td>
<td>3.6577</td>
<td>3.8842</td>
<td>4.4339</td>
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</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>N.A.</td>
<td>6.5402</td>
<td>4.7666</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(SSK, SM–QAM)</td>
<td>0.3574</td>
<td>0.2639</td>
<td>$-2.5664$</td>
<td>$-3.1516$</td>
<td>$-5.7457$</td>
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</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>0.1096</td>
<td>0.0934</td>
<td>$-2.6194$</td>
<td>$-2.9926$</td>
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</tr>
<tr>
<td></td>
<td>N.A.</td>
<td>N.A.</td>
<td>0.0401</td>
<td>0.0367</td>
<td>$-2.6598$</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 2. ABEP of QAM ($M_{QAM}^Q = 64$) and SM-QAM ($M = 32, N_t = 2$) against $E_m/N_0$. Accuracy of proposed analytical framework (denoted by “improved union-bound” in the legend) and conventional union bound (denoted by “union-bound” in the legend) for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 6$ bpcu).
Fig. 3. ABEP of SM-PSK against $E_m/N_0$. Performance comparison for various sizes of signal and spatial constellation diagrams. Accuracy of proposed analytical frameworks for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 4$ bpcu). The setup ($M = 2, N_t = 8$) is not shown as it overlaps with the setup ($M = 4, N_t = 4$).
Fig. 4. ABEP of SM-QAM against $E_m/N_0$. Performance comparison for various sizes of signal and spatial constellation diagrams. Accuracy of proposed analytical frameworks for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 4$ bpcu). The setup ($M = 2, N_t = 8$) is not shown as it overlaps with the setup ($M = 4, N_t = 4$).
Fig. 5. ABEP of SM-PSK against $E_m/N_0$. Performance comparison for various sizes of signal and spatial constellation diagrams. Accuracy of proposed analytical frameworks for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 5$ bpcu). The setup ($M = 2, N_t = 16$) is not shown as it overlaps with the setup ($M = 4, N_t = 8$).
Fig. 6. ABEP of SM-QAM against $E_m/N_0$. Performance comparison for various sizes of signal and spatial constellation diagrams. Accuracy of proposed analytical frameworks for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 5$ bpcu). The setup ($M = 2, N_t = 16$) is not shown as it overlaps with the setup ($M = 4, N_t = 8$).
Fig. 7. ABEP of SM-PSK against $E_m/N_0$. Performance comparison for various sizes of signal and spatial constellation diagrams. Accuracy of proposed analytical frameworks for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 6$ bpcu). The setup ($M = 2, N_t = 32$) is not shown as it overlaps with the setup ($M = 4, N_t = 16$).
Fig. 8. ABEP of SM-QAM against $E_m/N_0$. Performance comparison for various sizes of signal and spatial constellation diagrams. Accuracy of proposed analytical frameworks for unit-power ($\sigma_0^2 = 1$) i.i.d. Rayleigh fading (the rate is $R = 6$ bpcu). The setup ($M = 2, N_t = 32$) is not shown as it overlaps with the setup ($M = 4, N_t = 16$).
Fig. 9. ABEP against $E_m/N_0$ over i.i.d. Nakagami-$m$ fading ($m_{Nak} = 1.0$, i.e., Rayleigh, $N_r = 2$, and rate $R = 6$ bpcu). Performance comparison and accuracy of the analytical framework for SM-QAM, QAM, and SSK.
Fig. 10. ABEP against $E_m/N_0$ over i.i.d. Nakagami-$m$ fading ($m_{Nak} = 0.5$ and $m_{Nak} = 1.5$, $N_r = 2$, and rate $R = 6$ bpcu). Performance comparison and accuracy of the analytical framework for SM-QAM, QAM, and SSK.
Fig. 11. ABEP of SM-QAM against $E_m/N_0$ over correlated (at the transmitter) and identically distributed Nakagami-$m$ fading ($m_{Nak} = 0.5$ and $m_{Nak} = 1.5$, $N_r = 2$, and rate $R = 6$ bpcu). Performance comparison and accuracy of the analytical framework for $M = 2$ and $N_t = 32$. 
Fig. 12. ABEP of SM-QAM against $E_m/N_0$ over correlated (at the transmitter) and identically distributed Nakagami-$m$ fading ($m_{\text{Nak}} = 0.5$ and $m_{\text{Nak}} = 1.5$, $N_r = 2$, and rate $R = 6$ bpcu). Performance comparison and accuracy of the analytical framework for $M = 32$ and $N_t = 2$. 
Error Performance – Main Trends (37/38)

Fig. 13. ABEP of SM-QAM against $E_m/N_0$ over correlated (at the receiver) and identically distributed Nakagami-$m$ fading ($m_{Nak} = 0.5$ and $m_{Nak} = 1.5$, $N_r = 2$, and rate $R = 6$ bpcu). Performance comparison and accuracy of the analytical framework for $M = 2$ and $N_t = 32$. 
Fig. 14. ABEP of SM-QAM against $E_m/N_0$ over correlated (at the receiver) and identically distributed Nakagami-$m$ fading ($m_{Nak} = 0.5$ and $m_{Nak} = 1.5$, $N_r = 2$, and rate $R = 6$ bpcu). Performance comparison and accuracy of the analytical framework for $M = 32$ and $N_t = 2$. 

Error Performance – Main Trends (38/38)
Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
8. Imperfect Channel State Information at the Receiver
9. Multiple Access Interference
10. Energy Efficiency
11. Transmit-Diversity for SM
12. Spatially-Modulated Space-Time-Coded MIMO
13. Relay-Aided SM
14. SM in Heterogeneous Cellular Networks
15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
Achievable Capacity (1/5)

- **Receiver Diversity case**: $n_t = 1, n_r = n$
  \[ C = \log_2[1 + \rho \chi_{2n}^2] \]

- **Transmit Diversity case**: $n_t = n, n_r = 1$
  \[ C = \log_2[1 + (\rho / n_T) \chi_{2n}^2] \]

- **Combined Transmit and Receiver Diversity**: $n_t \geq n_r$
  \[ C > \sum_{k=n_T-(n_R-1)}^{n_T} \log_2[1 + (\rho / n_T) \chi_{2k}^2] \]

- **Cycling using one transmitted at a time**:
  \[ C = (1 / n_T) \sum_{i=1}^{n_T} \log_2[1 + \rho \chi_{2n_Ri}^2] \]

Achievable Capacity (2/5)

\[
C_{SM} (N_t \times 1) = C_1 + C_2 \approx C_1
\]

\[
C_1 = \frac{1}{N_t} \sum_{m=1}^{N_t} \log_2 \left( 1 + \rho |h_m|^2 \right)
\]

\[
C_2 = \frac{1}{N_t} \sum_{m=1}^{N_t} \left[ \int_y f(y| h_m) \log_2 \left( \frac{f(y| h_m)}{f(y)} \right) dy \right]
\]

\[
f(y) = \frac{1}{N_t} \sum_{m=1}^{N_t} \frac{1}{\pi \left( |h_m|^2 \sigma_X^2 + \sigma_N^2 \right)} \exp \left( - \frac{|y|^2}{|h_m|^2 \sigma_X^2 + \sigma_N^2} \right) f(y| h_m)
\]

Achievable Capacity (3/5)

Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
8. Imperfect Channel State Information at the Receiver
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13. Relay-Aided SM
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15. SM for Visible Light Communications
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18. Implementation Challenges of SM-MIMO
The performance of SSK/SM modulation significantly depends on the wireless channel statistics, and power imbalance may improve the performance.

Can power imbalance be created via opportunistic power allocation?

Assumptions:
- $N_t = 2$
- Correlated Rayleigh fading channel
- $E_1 \neq E_2$

$$\begin{align*}
\text{ABEP}(E_1, E_2) &= \frac{1}{2} \left[ 1 - \frac{1}{2} \sqrt{\frac{\sigma^2 \bar{\gamma}}{1 + \sigma^2 \bar{\gamma}}} \right] \\
\bar{\sigma}^2 &= E_1 \sigma_1^2 + E_2 \sigma_2^2 - 2 \rho \sqrt{E_1} \sqrt{E_2} \sigma_1 \sigma_2 \quad \text{and} \quad \bar{\gamma} = 1/(4N_0)
\end{align*}$$

Channel State Information at the Transmitter (2/22)

\[
\begin{align*}
\left( E_1^*, E_2^* \right) &= \arg \min \left\{ \text{ABEP} \left( E_1, E_2 \right) \right\} \\
\text{subject to: } & \frac{E_1}{2} + \frac{E_2}{2} = E_{\text{av}}
\end{align*}
\]

- If $\sigma_1^2 > \sigma_2^2 \rightarrow (E_1^*, E_2^*) = (2E_{\text{av}}, 0)$ and $\sigma_M^2 = \sigma_1^2$
- If $\sigma_2^2 > \sigma_1^2 \rightarrow (E_1^*, E_2^*) = (0, 2E_{\text{av}})$ and $\sigma_M^2 = \sigma_2^2$

\[
\text{SNR}_{\text{gain}} \approx 10 \log_{10} \left[ \frac{2\sigma_M^2}{\sigma_1^2 + \sigma_2^2 - 2\rho\sigma_1\sigma_2} \right] \geq 0
\]
Channel State Information at the Transmitter (3/22)
Channel State Information at the Transmitter (4/22)

OOSSK
The symbol error rate (SER) performance highly depends on the Euclidean distance between pairs of these vectors.

Optimization problem: how to design the transmit vectors using CSIT such that the distance between pairs of constellation vectors at the receiver is larger.

Two methods are proposed:
- In the first method, no constraint on the structure of the transmit vectors is imposed (Multi-Antenna Space Modulation: MSMod).
- In the second method, the transmit vectors have only one non-zero entry (Modified Space Shift Keying: MSSK).

Channel State Information at the Transmitter (6/22)

MSMod with Full-CSIT

\[
P_M \big| \mathbf{H} \leq \frac{1}{N_t} \sum_{k=1}^{N_t} Q \left( \sqrt{\frac{\rho}{2}} \| \mathbf{H} \mathbf{p}_l - \mathbf{H} \mathbf{p}_k \| \right)
\]

\[
\leq \frac{1}{2N_t} \sum_{k=1}^{N_t} \exp \left( -\gamma \| \mathbf{H} \mathbf{p}_l - \mathbf{H} \mathbf{p}_k \|^2 \right) = A_P
\]

\[
\begin{align*}
\left\{ (\mathbf{p}_1^*, \mathbf{p}_2^*, \ldots, \mathbf{p}_{N_t}^*) = \arg \min_{(\mathbf{p}_1, \mathbf{p}_2, \ldots, \mathbf{p}_{N_t})} \{ A_P \} \\
\text{s.t.} \quad \sum_{l=1}^{N_t} \| \mathbf{p}_l \|^2 = N_t
\end{align*}
\]

MSMod with Full-CSIT: Optimal Solution

\[ \mathbf{p}_k^* = \theta_k \mathbf{v}_1 \]

\[ \sum_{k=1, k \neq l}^{N_t} \left( 1 - \frac{\theta_k}{\theta_l} \right) \exp \left( -\gamma \varepsilon_1^2 |\theta_l - \theta_k|^2 \right) = \frac{\lambda'}{\varepsilon_1^2} \]

- \( \mathbf{v}_1 \) is the right singular vector related to the largest singular value of \( \mathbf{H} \)
- \( \varepsilon_1 \) is the largest singular value of \( \mathbf{H} \)
- \( \lambda' \) is a constant
- Bottom line: \( \theta_k \) can be chosen from conventional PSK/QAM constellations
- Similar results apply to the imperfect CSIT case (\( \mathbf{H}_{\text{error}} = \mathbf{H} + \mathbf{N} \))

MSSK with Full-CSIT

\[
P_M |H| \leq \frac{1}{N_t} \sum_{l=1, k=1}^{N_t} Q \left( \sqrt{\frac{\rho}{2}} \| p_l h_l - p_k h_k \| \right)
\]

\[
\leq \frac{1}{2N_t} \sum_{l=1, k=1}^{N_t} \exp \left( -\gamma \| p_l h_l - p_k h_k \|^2 \right) = A_P
\]

\[
\begin{align*}
(p_1^*, p_2^*, \ldots, p_{N_t}^*) &= \arg \min_{(p_1, p_2, \ldots, p_{N_t})} \{ A_P \} \\
\text{s.t.} \quad \sum_{l=1}^{N_t} |p_l|^2 &= N_t
\end{align*}
\]

MSSK with Full-CSIT: Optimal Solution

- Find

\[ p_i = r_i \exp(j\varphi_i) \]

- Such that the following function is MINIMIZED:

\[
\sum_{\substack{k=1, l=1 \\ k \neq l}}^{N_t} \exp \left\{ -\gamma \left( r_l^2 \|h_l\|^2 + r_k^2 \|h_k\|^2 - 2r_l r_k \omega_{l,k} \right) \right\}
\]

where \( \omega_{l,k} = \Re \left\{ e^{j(\varphi_k - \varphi_l)} h_l^H h_k \right\} \).

MSSK with Full-CSIT: Optimal Solution

If $N_t = 2$:

\[
\begin{align*}
    p_1 &= \sqrt{1 + \frac{\|h_1\|^2 - \|h_2\|^2}{\sqrt{4\mu^2 + (\|h_1\|^2 - \|h_2\|^2)^2}}} \cdot \exp\left(j(\pi + \phi_2 - \phi)\right) \\
    p_2 &= \sqrt{1 + \frac{\|h_2\|^2 - \|h_1\|^2}{\sqrt{4\mu^2 + (\|h_1\|^2 - \|h_2\|^2)^2}}} \cdot \exp\left(j\phi_2\right)
\end{align*}
\]

with

\[
    h_2^H h_1 = \mu \exp\left(j\phi\right)
\]

\[
    \phi_1 - \phi_2 = \pi - \phi
\]

MSSK with Full-CSIT: Optimal Solution

- If \( N_t > 2 \), a sub-optimal iterative approach is proposed:
  - In each iteration, the pair of vectors with \( s \)-th minimum distance is considered and the optimal solution for \( N_t = 2 \) is computed.
  - To guarantee that the error performance does not increase with the iterations, an error function is introduced:

\[
err(x, l) \triangleq \sum_{i=1, i \neq l}^{N_t} Q\left(\frac{\sqrt{\frac{\rho}{2}} \|x h_i - p_i h_i\|}{\sqrt{\frac{\rho}{2}} \|x h_l - p_l h_l\|}\right)
\]

- Iteration over \( s \): \( s \)-th minimum distance over pairs of transmission vectors

---

```
Algorithm 1 MSSK signal design algorithm with full CSIT \( (N_t > 2) \)
1. \( p_1 = p_2 = \ldots = p_{N_t} = 1, \ s = 1 \)
2. \( G = H \cdot \text{diag} \{p_1, p_2, \ldots, p_{N_t}\} \)
3. \( (l^*, k^*) = \arg \min_{l,k} \{||g_l - g_k||, s\} \)
4. \( m_{l^*,k^*} = \sqrt{\frac{||p_{l^*}||^2 + ||p_{k^*}||^2}{2}} \)
5. Calculate \( q_{l^*} \) and \( q_{k^*} \) using (36) with \( \varphi^* = 0 \)
6. \( \varphi^* = \arg \min \{err(q_{l^*}, \exp(j \varphi^*), l^*) + err(q_{k^*}, \exp(j \varphi^*), k^*)\} \)
7. \( q_{l^*} = q_{l^*} \cdot \exp(j \varphi^*), \ q_{k^*} = q_{k^*} \cdot \exp(j \varphi^*) \)
8. If \( err(q_{l^*}, l^*) + err(q_{k^*}, k^*) \leq err(p_{l^*}, l^*) + err(p_{k^*}, k^*) \)
   - \( s = 1 \)
   - \( p_{l^*} = q_{l^*} \)
   - \( p_{k^*} = q_{k^*} \)
   else
   - \( s = s + 1 \)
9. Go to step 2
```
Fig. 2. SER for SSK, MSSK, and MSMod schemes with full CSIT ($N_t = 2$, $N_r = 2, 4$)
Fig. 3. SER for SSK, MSSK, and MSMod schemes with full CSIT ($N_t = 4$, $N_r = 2, 4$)
Fig. 4. SER for SSK, MSSK, and MSMod schemes with imperfect CSIT 
($N_t = 2$, $N_r = 2$, $\sigma_N^2 = 1$)
Fig. 5. Performance comparison of MSMod with other MIMO transmission schemes with CSIT ($N_t = 4$, $N_r = 4$)
Fig. 6. Performance comparison of modified SSK/SM with antenna selection.
The Approach

\[ P(x_i \rightarrow x_j | H) \approx \lambda \cdot Q \left( \sqrt{\frac{1}{2N_0} d_{\text{min}}^2(H)} \right) \]

\[ d_{\text{min}}(H) = \max_{x_i, x_j \in \Lambda, x_i \neq x_j} \| H(x_i - x_j) \|_F \]

The Proposed Adaptive Transmission Schemes

\[ [\tilde{q}, \tilde{d}] = \arg \max_{q_i \in \{K_i, \ldots, \tilde{q}_i\}} \min_{d_j \in \{d_j^1, \ldots, d_j^K\}} d_{\text{min}}(q_i, d_j) \]

subject to

\[
\begin{align*}
\left\{ q_i, d_j \mid \log(d_i \cdot K_i) = m, \text{ and } d_j^k = d_i \right\} & \quad \text{for AMS-SM} \\
\left\{ q_i, d_j \mid \log(K) + \frac{1}{K} \log \left( \prod_{k=1}^{K} d_j^k \right) = m, \text{ and } K \in \{K_1, \ldots, K_L\} \right\} & \quad \text{for ASM} \\
\left\{ q_i, d_j \mid \log(K_i) + \frac{1}{K_i} \log \left( \prod_{k=1}^{K_i} d_j^k \right) = m, \text{ and } i = 1, \ldots, L \right\} & \quad \text{for OH-SM} \\
\left\{ q_i, d_j \mid \log(K_{\text{AMS}}) + \frac{1}{K_{\text{AMS}}} \log \left( \prod_{k=1}^{K_{\text{AMS}}} d_j^k \right) = m, \text{ and } \tilde{q}_i = q_{\text{AMS}} \right\} & \quad \text{for C-SM}
\end{align*}
\]

- **AMS-SM**: Adaptive Modulation Scheme Spatial Modulation
- **ASM**: Adaptive Spatial Modulation
- **OH-SM**: Optimal Hybrid Spatial Modulation
- **C-SM**: Concatenated Spatial Modulation

Fig. 2. BER performance at 3 bits/s/Hz for link-adaptation and conventional SM schemes under uncorrelated channel conditions.
Fig. 3. BER performance at 4 bits/s/Hz for link-adaptation and conventional SM schemes under uncorrelated channel conditions.
Fig. 5. BER performance at 3 bits/s/Hz for link-adaptation and conventional SM schemes under correlated channel conditions with $r = 0.5$ and 0.9.
Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
8. Imperfect Channel State Information at the Receiver
9. Multiple Access Interference
10. Energy Efficiency
11. Transmit-Diversity for SM
12. Spatially-Modulated Space-Time-Coded MIMO
13. Relay-Aided SM
14. SM in Heterogeneous Cellular Networks
15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
The working principle of SM/SSK is based on the following facts:

1. The wireless environment naturally modulates the transmitted signal
2. Each transmit-receive wireless link has a different channel
3. The receiver employs the a priori channel knowledge to detect the transmitted signal
4. Thus, part of the information is conveyed by the Channel Impulse Response (CIR), i.e., the channel/spatial signature

How Much Important is Channel State Information for SSK/SM Modulation?
Imperfect Channel State Information at the Receiver (2/23)

- **Perfect CSI (channel gains and phases):** F–CSI (SSK)

\[
\hat{m} = \arg \max_{\{m_i\}_{i=1}^{N_t}} \{D_i\} \quad D_i = \text{Re} \left\{ \int_{T_m} r(t) \overline{s_i^*(t)} \, dt \right\} - \frac{1}{2} \int_{T_m} \overline{s_i(t)} \overline{s_i^*(t)} \, dt
\]

- **Partial CSI (channel gains):** P–CSI (SSK)

\[
\hat{m} = \arg \max_{\{m_i\}_{i=1}^{N_t}} \{\ln[D_i]\} \quad \overline{D}_i = \ln[D_i] = \begin{cases} 
\ln \left[ I_0 \left( \frac{\sqrt{E_m} \beta_i}{N_0} |\overline{r}| \right) \right] - \frac{E_m \beta_i^2}{2N_0} \\
\sqrt{E_m} \beta_i |\overline{r}| - \frac{E_m \beta_i^2}{2}
\end{cases}
\]

Imperfect Channel State Information at the Receiver (3/23)

\[
\begin{align*}
\begin{cases}
 r(t) &= \sqrt{E_m} \beta_i \exp(j \varphi_t) w(t) + n(t) \\
 \bar{r} &= \int_{T_m} r(t) w^*(t) \, dt
\end{cases}
\end{align*}
\]

\[
\begin{align*}
\hat{m} &= \arg \max \{ \ln[D_i] \} \\
\bar{D}_i &= \ln[D_i] = \sqrt{E_m} \beta_i |\bar{r}| - \frac{E_m \beta_i^2}{2}
\end{align*}
\]

\[
\begin{align*}
\text{ABEP} &\leq \frac{2}{N_t \log_2 (N_t)} \sum_{i_1}^{N_t} \sum_{i_2 = i_1+1}^{N_t} N(i_1, i_2) \text{AEP}(\text{TX}_{i_2} \rightarrow \text{TX}_{i_1}) \\
\text{AEP} &= \mathbb{E}_{h_{i_1}, h_{i_2}} \{ P_{E}(1, 2) \}
\end{align*}
\]

\[
P_{E}(1, 2) = \left[ \frac{1}{2} - \frac{1}{2} Q \left( \frac{\sqrt{E_m} \beta_1}{\sqrt{N_0}}, \frac{\sqrt{E_m} (\beta_1 + \beta_2)}{2\sqrt{N_0}} \right) \right] \Pr\{\beta_1 \geq \beta_2\}
\]

\[
+ \left[ \frac{1}{2} - \frac{1}{2} Q \left( \frac{\sqrt{E_m} \beta_2}{\sqrt{N_0}}, \frac{\sqrt{E_m} (\beta_1 + \beta_2)}{2\sqrt{N_0}} \right) \right] \Pr\{\beta_1 < \beta_2\}
\]

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2x1 MIMO, Correlated ($q=0.64$) Nakagami-m Fading

Scenario a:
$\Omega_1=1$, $\Omega_2=1$, $m_1=2$, $m_2=5$

Scenario b:
$\Omega_1=10$, $\Omega_2=1$, $m_1=2$, $m_2=5$

Scenario c:
$\Omega_1=10$, $\Omega_2=1$, $m_1=5$, $m_2=2$

Imperfect Channel State Information at the Receiver (5/23)

4x1 MIMO, Correlated (exponential) Nakagami-m Fading

Balanced:
\[ \{ \Omega_i \}_{i=1, \ldots, 4} = 1 \]

Unbalanced:
\[ \Omega_1 = 1, \{ \Omega_i \}_{i=2, \ldots, 4} = 4i-4 \]

Correlation:
\[ \rho_{i,j} = \exp(-d_0 |i-j|) \]
\[ d_0 = 0.22 \]

2x1 MIMO, Uncorrelated Nakagami-m Fading

Imperfect Channel State Information at the Receiver (7/23)

4x1 MIMO, Correlated (exponential) Nakagami-m Fading

SSK with Mismatched Decoder

Received Signal \rightarrow \text{SSK ML-Detector} \rightarrow \text{Estimated Antenna Index}

\hat{\alpha} = \alpha + \epsilon

\begin{align*}
\hat{m} &= \arg\min_{m_t \text{ for } t=1,2,\ldots,N_t} \left\{ \hat{D}_{m_q} (m_t) \right\} \\
&= \arg\min_{m_t \text{ for } t=1,2,\ldots,N_t} \left\{ \sum_{r=1}^{N_r} \left[ \frac{\tilde{\eta}_{0,r}}{\sqrt{N_0}} - \sqrt{\frac{E_m}{N_0}} (\alpha_{t,r} - \alpha_{q,r}) + \sqrt{\frac{E_m}{N_0}} \epsilon_{t,r} \right]^2 \right\}
\end{align*}

Imperfect Channel State Information at the Receiver (9/23)

ABEP

\[
\text{ABEP} \leq \frac{1}{N_t \log_2(N_t)} \sum_{q=1}^{N_t} \sum_{t=1}^{N_t} \left[ N_H \left( t, q \right) \left( \frac{1}{2} - \frac{1}{\pi} \int_0^{+\infty} \left[ \frac{\text{Im} \left\{ \gamma \left( \nu \right) M_{\gamma_t,q} \left( \Delta \left( \nu \right) \right) \right\}}{\nu} \right) d\nu \right) \right]
\]

\[
M_{\gamma_{t,q}} \left( s \right) = \mathbb{E} \{ \exp \left( -s \gamma_{t,q} \right) \} \quad \gamma_{t,q} = \gamma \left( \alpha \right) = \sum_{r=1}^{N_r} \left| \alpha_{q,r} - \alpha_{t,r} \right|^2
\]

Methodology for computation:

1. Union bound: the ABEP can be obtained from the APEP

\[
\text{APEP} \left( m_q \rightarrow m_t \right) = \mathbb{E} \left\{ \text{Pr} \left\{ \hat{D}_{m_q} \left( m_t \right) < \hat{D}_{m_q} \left( m_q \right) \right\} \right\} = \text{Pr} \left\{ \hat{D}_{m_q} \left( m_t \right) - \hat{D}_{m_q} \left( m_q \right) < 0 \right\}
\]

2. The (difference) decision variable is a quadratic-form in complex Gaussian RVs (when conditioning upon fading channel statistics)

3. The PEP is obtained by using the Gil-Pelaez inversion theorem

If $w_1(t) = w_2(t) \rightarrow$ Diversity $= N_r$ (conventional SSK)

If $w_1(t)$ is “time-orthogonal” to $w_2(t) \rightarrow$ Diversity $= 2N_r$ (TOSD-SSK)

This is true for any $N_t$ with no bandwidth expansion and with a single active transmit-antenna at any time-instance

TOSD-SSK with Mismatched Decoder

Received Signal ➔ TOSD-SSK ML-Detector ➔ Estimated Antenna Index

\[ \hat{\alpha} = \alpha + \varepsilon \]

ML Channel Estimator

\[
\hat{m} = \arg \min_{m_t \text{ for } t=1,2,...,N_t} \left\{ \hat{D}_{m_q} \left( m_t \right) \right\}
\]

\[
= \arg \min_{m_t \text{ for } t=1,2,...,N_t} \left\{ \sum_{r=1}^{N_r} \text{Re} \left\{ \alpha_{q,r} \hat{\alpha}_{t,r}^* E_m \delta_{t,q} + \hat{\alpha}_{t,r}^* \sqrt{E_m} \tilde{\eta}_{t,r} \right\} - \frac{E_m}{2} \sum_{r=1}^{N_r} |\hat{\alpha}_{t,r}|^2 \right\}
\]

Perfect Channel State Information at the Receiver (12/23)

Methodology for computation:

1. Union bound: the ABEP can be obtained from the APEP

\[
\text{APEP}(m_q \to m_t) = \mathbb{E}\left\{ \Pr \left\{ \hat{D}_{m_q}(m_t) < \hat{D}_{m_q}(m_q) \right\} \right\} = \Pr \left\{ \hat{D}_{m_q}(m_t) - \hat{D}_{m_q}(m_q) < 0 \right\}
\]

2. The (difference) decision variable is the difference of two independent quadratic-forms in complex Gaussian RVs (when conditioning upon fading channel statistics)

3. The PEP is obtained by using the Gil-Pelaez inversion theorem

Diversity Analysis (i.i.d. Rayleigh Fading)

**SSK**

\[
\text{ABEP} \leq \frac{N_t}{4} - \frac{N_t}{2\pi} \int_{0}^{+\infty} \left[ \text{Im} \left\{ \frac{\Upsilon (\nu)}{\nu} \frac{1}{(1 - 2\Omega_0 \Delta (\nu))^{Nr}} \right\} \right] d\nu 
\]

**TOSD-SSK**

\[
\text{ABEP} \leq \frac{N_t}{4} - \frac{N_t}{2\pi} \int_{0}^{+\infty} \left[ \text{Im} \left\{ \frac{\Upsilon_q (\nu) \Upsilon_t (-\nu)}{\nu} \frac{1}{(1 - \Omega_0 \Delta_q (\nu))^{Nr} (1 - \Omega_0 \Delta_t (-\nu))^{Nr}} \right\} \right] d\nu 
\]

**With channel estimation errors:**

1. Diversity order of SSK is: \(Nr\)
2. Diversity order of TOSD-SSK is: \(2Nr\)

Numerical Results (SSK)
Numerical Results (TOSD-SSK)
Imperfect Channel State Information at the Receiver (16/23)

Single-Antenna MQAM

![Graph showing ABEP vs. E_m/N_0 for different values of N_r and N_p. The graph illustrates the performance of MQAM with varying channel conditions.](image)
Imperfect Channel State Information at the Receiver (17/23)

Alamouti MQAM

![Graph](image)
## Imperfect Channel State Information at the Receiver (18/23)

### SSK vs. Single-Antenna MQAM (Nr=1 / Nr=2 / Nr=4)

<table>
<thead>
<tr>
<th>Rate</th>
<th>$N_p = 1$</th>
<th>$N_p = 3$</th>
<th>$N_p = 10$</th>
<th>P – CSI</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 bpcu</td>
<td>22.9 / 25.3 / 16.2</td>
<td>21.1 / 23.5 / 14.5</td>
<td>20.3 / 22.7 / 13.6</td>
<td>19.9 / 22.3 / 13.2</td>
</tr>
<tr>
<td>3 bpcu</td>
<td>29 / 28.4 / 17.9</td>
<td>27.3 / 26.6 / 16.2</td>
<td>26.4 / 25.8 / 15.3</td>
<td>26 / 25.4 / 14.9</td>
</tr>
<tr>
<td>4 bpcu</td>
<td>32 / 29.9 / 18.7</td>
<td>30.3 / 28.1 / 17</td>
<td>29.5 / 27.3 / 16.2</td>
<td>29 / 26.9 / 15.7</td>
</tr>
</tbody>
</table>

### Single-Antenna QAM

<table>
<thead>
<tr>
<th>Rate</th>
<th>$N_p = 1$</th>
<th>$N_p = 3$</th>
<th>$N_p = 10$</th>
<th>P – CSI</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 bpcu</td>
<td>19.8 / 22.3 / 13.1</td>
<td>18.2 / 20.6 / 11.5</td>
<td>17.2 / 19.7 / 10.6</td>
<td>16.8 / 19.3 / 10.1</td>
</tr>
<tr>
<td>2 bpcu</td>
<td>22.7 / 25.2 / 16.2</td>
<td>21.1 / 23.5 / 14.5</td>
<td>20.3 / 22.7 / 13.6</td>
<td>19.9 / 22.2 / 13.2</td>
</tr>
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<td>25.6 / 28 / 19.1</td>
<td>24.9 / 27.4 / 18.2</td>
<td>24.6 / 27 / 18.1</td>
</tr>
<tr>
<td>4 bpcu</td>
<td>29.6 / 32 / 23.3</td>
<td>27.8 / 30.1 / 21.3</td>
<td>27.1 / 29.6 / 20.5</td>
<td>26.8 / 29.1 / 20.3</td>
</tr>
</tbody>
</table>

**Take Away Message:**
- SSK is better than single-antenna MQAM if Rate>2bpcu and Nr>1
- The robustness to channel estimation errors is the same
**Imperfect Channel State Information at the Receiver (19/23)**

**TOSD-SSK vs. Alamouti MQAM** (Nr=1 / Nr=2)

**Take Away Message:**
- TOSD-SSK is better than Alamouti MQAM if Rate > 2 bpcu
- TOSD-SSK is more robust to channel estimation errors

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</tr>
<tr>
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<td>27.5 / 17.8</td>
<td>27 / 17.3</td>
<td>26.8 / 17</td>
</tr>
<tr>
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<td>25.4 / 16.3</td>
</tr>
<tr>
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<td>31.4 / 22.2</td>
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</tr>
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<td>4 bpcu</td>
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<td>33.3 / 24.3</td>
<td>32.6 / 23.5</td>
<td>32.3 / 23.3</td>
</tr>
</tbody>
</table>
SM with Imperfect CSIR

- Channel estimation model:
  \[
  \begin{cases}
  \beta = \alpha + \varepsilon \quad \text{with} \quad \rho(\alpha, \beta) = 1/\sqrt{1 + \sigma^2_{\varepsilon}} \\
  \sigma^2_{\varepsilon} = \text{const} \quad \text{and} \quad \sigma^2_{\varepsilon} = 1/(\gamma N)
  \end{cases}
  \]

- SM with MPSK modulation:
  \[
  (\hat{j}, \hat{s}) = \arg\min_{j, s} \left\{ \sum_{r=1}^{N_r} |y_r - \rho^2 \beta_{j,r,s}|^2 \right\}
  \]

- SM with MQAM modulation:
  \[
  (\hat{j}, \hat{s}) = \arg\min_{j, s} \left\{ \sum_{r=1}^{N_r} |y_r - \beta_{j,r,s}|^2 \right\}
  \]

Fig. 1. BER performance of SM with $n_T = 4$, QPSK and V-BLAST with $n_T = 4$, BPSK (4 bits/s/Hz) with optimal receivers (fixed $\sigma_e^2$).

Fig. 2. BER performance of SM with $n_T = 4$, QPSK and V-BLAST with $n_T = 4$, BPSK (4 bits/s/Hz) with optimal receivers (variable $\sigma_e^2$).

Fig. 3. BER performance of SM with $n_T = 4$, 16-QAM and V-BLAST with $n_T = 3$, QPSK (6 bits/s/Hz) with mismatched receivers (fixed $\sigma_e^2$).

Outline

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2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
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5. Error Performance (Numerical Results and Main Trends)
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15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
The working principle of SM/SSK is based on the following facts:

1. The wireless environment naturally modulates the transmitted signal
2. Each transmit-receive wireless link has a different channel
3. The receiver employs the a priori channel knowledge to detect the transmitted signal
4. Thus, part of the information is conveyed by the Channel Impulse Response (CIR), i.e., the channel/spatial signature

Can the randomness of the fading channel be used for Multiple-Access too rather than just for Modulation?
Multiple Access Interference (2/22)

Signal Model

\[ z^{(r)}(t) = \sqrt{E_{\xi}} h_{\xi}^{(x_{\xi},r)} w(t) + \sum_{u=1}^{N_u} \left[ \sqrt{E_{u}} h_{u}^{(x_{u},r)} w(t) \right] + n^{(r)}(t) \]

- probe link
- interference
- AWGN

Single-User Detector

\[ \hat{x}_{\xi} = \arg \min_{y_{\xi}=1,2,...,N_t} \left\{ D^{(y_{\xi})} \right\} = \arg \min_{y_{\xi}=1,2,...,N_t} \left\{ \sum_{r=1}^{N_r} \int_{T_s} \left| z^{(r)}(t) - \sqrt{E_{\xi}} h_{\xi}^{(y_{\xi},r)} w(t) \right|^2 dt \right\} \]

Multi-User Detector

\[ \hat{x} = \arg \min_{y=[y_1,y_2,...,y_{N_t}]} \left\{ D^{(y)} \right\} = \arg \min_{y=[y_1,y_2,...,y_{N_t}]} \left\{ \sum_{r=1}^{N_r} \int_{T_s} \left| z^{(r)}(t) - \sum_{u=1}^{N_u} \sqrt{E_{u}} h_{u}^{(y_{u},r)} w(t) \right|^2 dt \right\} \]

Multiple Access Interference (3/22)

SSK with Single-User Detector (i.i.d. Rayleigh)

\[
\text{ABEP}_\xi \leq \frac{N_t}{2} \left[ \frac{1}{2} \left( 1 - \sqrt{\frac{\text{SINR}}{2 + \text{SINR}}} \right) \right]^{N_r} \sum_{r=1}^{N_r} \left\{ \left( N_r - 1 + r \right) \left[ \frac{1}{2} \left( 1 + \sqrt{\frac{\text{SINR}}{2 + \text{SINR}}} \right) \right]^r \right\}
\]

\[
\begin{align*}
\text{SINR} &= \text{SNR}_\xi / (1 + \text{INR}_\setminus \xi) \\
\text{SNR}_\xi &= \left( E_\xi \sigma_\xi^2 \right) / N_0 \quad \text{and} \quad \text{INR}_\setminus \xi = \sum_{u \neq \xi = 1}^{N_u} \left[ \left( E_u \sigma_u^2 \right) / N_0 \right]
\end{align*}
\]

- \( E_u = 0 \) (no interference): framework reduces to single-user case
- \( \text{SNR}_\xi \gg 1 \) and \( \text{INR}_\setminus \xi \ll 1 \) (noise limited):

\[
\text{ABEP}_\xi \rightarrow 2^{-(N_r + 1)} \left( 2^{N_r} - 1 \right) N_t \text{SNR}_\xi^{-N_r}
\]

SSK with Single-User Detector (i.i.d. Rayleigh)

- $\text{INR}_\xi \gg 1$ and $\text{SIR} = \text{SNR}_\xi / \text{INR}_\xi$ ($\gg 1$) (interference limited):

$$\text{ABEP}_\xi \to 2^{-(N_r+1)} \left(\frac{2N_r}{N_r} - 1 \right) N_t \text{SIR}^{-N_r} \text{ with SIR} = E_\xi \sigma_\xi^2 / \sum_{u \neq \xi}^{N_u} (E_u \sigma_u^2)$$

- $N_r \gg 1$:

$$\text{ABEP}_\xi \to \left(\frac{N_t}{2}\right) Q\left(\sqrt{N_r \text{SINR}}\right)$$

$$\begin{cases} 
\text{SINR} = \frac{\text{SNR}_\xi}{(1 + \text{INR}_\xi)} \\
\text{SNR}_\xi = \frac{(E_\xi \sigma_\xi^2)}{N_0} \quad \text{and} \quad \text{INR}_\xi = \sum_{u \neq \xi}^{N_u} \left[\frac{(E_u \sigma_u^2)}{N_0}\right]
\end{cases}$$

Multiple Access Interference (5/22)

SSK vs. MPSK/MQAM (Single-User Detector, i.i.d. Rayleigh)

\[ M = N_t \text{ (same bpcu)} \]

\[ \frac{\text{ABEP}_{\text{PSK}}^{\xi}}{\text{ABEP}_{\text{SSK}}^{\xi}} \rightarrow \frac{2}{N_t^2 \log_2 (N_t)} \sum_{x_{\xi}=1}^{N_t} \sum_{y_{\xi}=1}^{N_t} \left\{ N_H \left( S_{\xi}(x_{\xi}), S_{\xi}(y_{\xi}) \right) \sqrt{\frac{2}{Q} \left( S_{\xi}(y_{\xi}) - S_{\xi}(x_{\xi}) \right)^2} \right\}^{N_r} \]

- SSK will never be better than MPSK/MQAM if \( Q \geq 2 \). This occurs if \( M = N_t = 2 \) and \( M = N_t = 4 \). If \( M = N_t > 4 \) a crossing point exists

- If \( Q < 2 \), the performance gain of SSK exponentially increases with \( N_r \)

GSSK with Single-User Detector (i.i.d. Rayleigh)

\[
\text{SINR} = \frac{1}{2} \frac{\text{SNR}_\xi}{1 + \text{INR}_\xi} \frac{N^{\neq}_{\text{ta}}(x_\xi, y_\xi)}{N_{\text{ta}}}
\]

- \(N_{\text{ta}}\) is the number of active antennas
- \(N_{\text{ta}}^{\neq}\) is the number of different antenna indexes: \(2 \leq N_{\text{ta}}^{\neq} \leq 2N_{\text{ta}}\)
- Asymptotic performance:

\[
\begin{cases}
\text{APEP}(x_\xi \rightarrow y_\xi) \rightarrow \left[N_{\text{ta}}/N_{\text{ta}}^{\neq}(x_\xi, y_\xi)\right]^{N_r} \left(\frac{2N_r}{N_r} - 1\right) \gamma^{-N_r} \\
\gamma = \text{SNR}_\xi (\text{noise limited}) \text{ or } \gamma = \text{SIR} (\text{interference limited})
\end{cases}
\]

Multiple Access Interference (7/22)

SSK vs. GSSK (Single-User Detector, i.i.d. Rayleigh)

\[
\frac{\text{APEP}^{\text{GSSK}}_{\xi} (x_\xi \rightarrow y_\xi)}{\text{APEP}^{\text{SSK}}_{\xi}} \rightarrow \left[ \frac{2N_{\text{ta}}}{N_{\text{ta}}^\# (x_\xi, y_\xi)} \right]^{N_r}
\]

- Since \(2 \leq N_{\text{ta}}^\# \leq 2N_{\text{ta}}\), GSSK is worse than SSK regardless of the choice of the spatial-constellation diagram.
- The SNR gap is:

\[
0 \leq \Delta_Y \leq 10 \log_{10} (N_{\text{ta}})
\]

thus, the larger \(N_{\text{ta}}\), the worse GSSK compared to SSK.

Multiple Access Interference (8/22)

SSK and GSSK with Multi-User Detector (i.i.d. Rayleigh)

\[
\text{APEP}(x \rightarrow y) = \left[ \frac{1}{2} \left( 1 - \sqrt{\frac{\text{AggrSNR}}{2 + \text{AggrSNR}}} \right) \right]^{N_r} \sum_{r=1}^{N_r} \left\{ (N_r - 1 + r) \left[ \frac{1}{2} \left( 1 + \sqrt{\frac{\text{AggrSNR}}{2 + \text{AggrSNR}}} \right) \right]^r \right\}
\]

- **SSK**  \( \text{AggrSNR} = \sum_{u=1}^{N_u} \left[ \frac{E_u \sigma_u^2 \left( 1 - \delta_{x_u, y_u} \right)}{N_0} \right] \)

- **GSSK**  \( \text{AggrSNR} = \sum_{u=1}^{N_u} \left[ \frac{E_u \sigma_u^2 N^{\#}_{ta}(x_u, y_u)}{2N_0} \frac{N_{ta}}{N_{ta}} \left( 1 - \delta_{x_u, y_u} \right) \right] \)

- Unlike the single-user detector, \( \text{APEP} \rightarrow 0 \) if \( N_0 \rightarrow 0 \)

SSK with Multi-User Detector (i.i.d. Rayleigh) – Asymptotic Analysis

- **AggSNR >> 1**

\[
\text{ABEP}_\xi \rightarrow \left[N_t^{N_u} \log_2 (N_t)\right]^{-1} \left(\frac{2N_r}{N_r} - 1\right) 2^{-N_r} \sum_x \sum_y \left[(1 - \delta_{x_\xi, y_\xi}) N_H(x_\xi, y_\xi) \text{AggrSNR}^{-N_r}\right]
\]

- **Single-user lower bound \((N_u = 1)\)**

\[
\text{ABEP}^\text{SULB}_\xi \rightarrow 2^{-(N_r+1)} \left(\frac{2N_r}{N_r} - 1\right) N_t \text{SNR}^{-N_r}
\]

- **SNR gap due to multiple-access interference**

\[
\Delta_{\text{SNR}} = \frac{10}{N_r} \log_{10} \left(\frac{\text{ABEP}_\xi}{\text{ABEP}^\text{SULB}_\xi}\right) \rightarrow 10 \log_{10} \left(\frac{1}{N_r(N_u+1)} \sum_x \sum_y \left(\frac{1 - \delta_{x_\xi, y_\xi}}{N_H(x_\xi, y_\xi) E_\xi \sigma^2}\right) \sum_{u=1}^{N_u} \left[E_u \sigma^2 u (1 - \delta_{x_u, y_u})\right]\right]
\]

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Multiple Access Interference (10/22)

SSK with Multi-User Detector (i.i.d. Rayleigh) – Asymptotic Analysis

- **Strong interference case** ($E_w \sigma_w^2 << E_u \sigma_u^2$, for every $u$)

  \[
  \text{ABEP}_w \rightarrow 2^{-(N_r+1)} N_t \left( \frac{2N_r}{N_r} - 1 \right) \left( E_w \sigma_w^2 / N_0 \right)^{-N_r} = \text{ABEP}^{\text{SULB}}_w
  \]

- **Weak interference case** ($E_b \sigma_b^2 >> E_u \sigma_u^2$, for every $u$)

  \[
  \text{ABEP}_b \rightarrow 2^{-(N_r+1)} N_t^{N_u} \left( \frac{2N_r}{N_r} - 1 \right) \left[ E_b \sigma_b^2 / N_0 \right]^{-N_r}
  \]

  \[
  \Delta_{\text{SNR}} = \left( 10 / N_r \right) \log_{10} \left( \frac{\text{ABEP}_b}{\text{ABEP}^{\text{SULB}}_b} \right) = 10 \left[ \left( N_u - 1 \right) / N_r \right] \log_{10} \left( N_t \right)
  \]

Multiple Access Interference (11/22)

SSK with Multi-User Detector (i.i.d. Rayleigh) – Asymptotic Analysis

- Generic user

\[ \text{ABEP}_u^L \leq \text{ABEP}_u \leq \text{ABEP}_u^U \]

\[
\begin{align*}
\text{ABEP}_u^L &= 2^{-(N_r+1)} N_t \left( \frac{2N_r - 1}{N_r} \right) \left( \frac{E_u \sigma_u^2}{N_0} \right)^{-N_r} \\
\text{ABEP}_u^U &= 2^{-(N_r+1)} N_t^{N_u} \left( \frac{2N_r - 1}{N_r} \right) \left[ \frac{E_u \sigma_u^2}{N_0} \right]^{-N_r}
\end{align*}
\]

\[
\Delta_{\text{SNR}_u} = \left( \frac{10}{N_r} \right) \log_{10} \left( \frac{\text{ABEP}_u^U}{\text{ABEP}_u^L} \right) = 10 \left[ (N_u - 1)/N_r \right] \log_{10} (N_t)
\]

**Multiple Access Interference (12/22)**

GSSK with Multi-User Detector (i.i.d. Rayleigh) – Asymptotic Analysis

\[
\begin{aligned}
\text{ABEP}^L_u \leftrightarrow \text{SULB} \quad \text{and} \quad \text{ABEP}^U_u \leftrightarrow \text{weak interference case} \\
\text{ABEP}^L_u \leq \text{ABEP}^u \leq \text{ABEP}^U_u \\
\begin{aligned}
2^{-\left(N_r+1\right)} N_t \left( \frac{2N_r}{N_r} - 1 \right) \left( \frac{E_u \sigma^2}{N_0} \right)^{-N_r} & \leq \text{ABEP}^L_u \leq 2^{-\left(N_r+1\right)} 2^{\left[ \log_2 \left( \frac{N_u}{N_{ta}} \right) \right]} N_{ta} \left( \frac{2N_r}{N_r} - 1 \right) \left( \frac{E_u \sigma^2}{N_0} \right)^{-N_r} \\
\end{aligned}
\end{aligned}
\]

\[
\Delta_{\text{SNR}} = \frac{\left(10/N_r\right) \log_{10} \left( \frac{\text{ABEP}^U_u}{\text{ABEP}^L_u} \right)}{N_t}
\]

Fig. 1. ABEP of SSK modulation with single-user detection. Setup: $N_t = 8$; $N_u = 2$; $\sigma_1^2 = 1$; and (left) $N_r = 1$ and (right) $N_r = 3$. Markers show Monte Carlo simulations, and solid lines show the analytical model [i.e., (8) and (9)]. The ABEP of user 1 (probe/intended link) is shown. SULB stands for SULB, i.e., it represents the scenario with no multiple-access interference.
Fig. 2. ABEP of SSK modulation with single-user detection. Setup: $N_t = 8$; $\sigma_i^2 = 1$; $\sigma_i^2 = 10^{-2}$ for $i = 2, 3, \ldots, N_u$; $N_r = 1$; and $N_r = 3$. Markers show Monte Carlo simulations, and solid lines show the analytical model [i.e., (8) and (9)]. The ABEP of user 1 (probe/intended link) is shown.
Fig. 3. ABEP of (left) SSK and (right) GSSK modulations with single-user detection. Setup: \( N_u = 3; \sigma_i^2 = 1; \sigma_i^2 = 10^{-2} \) for \( i = 2, 3, \ldots, N_u; \) and \( N_r = 2 \). Markers show Monte Carlo simulations, and solid lines show the analytical model (i.e., (8) and (9) for SSK modulation and (13) for GSSK modulation). The ABEP of user 1 (probe/intended link) is shown.

Fig. 4. ABEP of (left) PSK and (right) QAM modulations with single-user detection. Setup: \( N_u = 3; \sigma_i^2 = 1; \sigma_i^2 = 10^{-2} \) for \( i = 2, 3, \ldots, N_u; \) and \( N_r = 2 \). Markers show Monte Carlo simulations, and solid lines show the analytical model (i.e., the union bound in the first and second rows of Table I). The ABEP of user 1 (probe/intended link) is shown.
Fig. 5. ABEP of (blue curves) SSK and (green curves) QAM modulations with single-user detection. Setup: (left) $N_u = 1$ and (right) $N_u = 2$; $\sigma^2 = 1$ and $\sigma^2 = 5 \times 10^{-2}$; and $N_r = 3$. Markers show Monte Carlo simulations, and solid lines show the analytical model (i.e., (8) and (9) for SSK modulation and the union bound in the second row of Table I). The ABEP of user 1 (probe/intended link) is shown.
Fig. 7. ABEP of, on the left, (blue curves) SSK and (green curves) QAM, and, on the right, (red curves) PSK and (magenta curves) GSSK modulations with single-user detection. Setup: $N_u = 2$; $\sigma_1^2 = 1$ and $\sigma_2^2 = 10^{-2}$; and $N_r = 3$. For GSSK modulation, we have $(N_t, N_{ta}) = (5, 2)$ if $N = 8$; $(N_t, N_{ta}) = (8, 4)$ if $N = 64$; $(N_t, N_{ta}) = (11, 4)$ if $N = 256$; and $(N_t, N_{ta}) = (12, 5)$ if $N = 1024$. For SSK and GSSK modulations, markers show Monte Carlo
Fig. 10. ABEP of SSK modulation with multiuser detection. Setup: $N_t = 8$; $N_u = 3$; $\sigma_1^2 = 0.1$, $\sigma_2^2 = 1$, $\sigma_3^2 = 10$; and $N_r = 3$. Markers show Monte Carlo simulations, and solid lines show the analytical model [i.e., (16) and (18)]. Furthermore, dashed lines show the estimated lower bound [i.e., $\text{ABEP}_{\text{L}}^u$ in (22)], which corresponds to the SULB when no multiple-access interference is present; and dotted lines show the estimated upper bound [i.e., $\text{ABEP}_{\text{U}}^u$ in (22)]. The ABEP of all the users is shown.
Multiple Access Interference (19/22)

Fig. 11. ABEP of SSK modulation with multiuser detection. Setup: $N_t = 8$; $\sigma_i^2 = 1$ and $\sigma_i^2 = 10^{-2}$ for $i = 2, 3, \ldots, N_u$; $N_r = 1$; and $N_r = 3$. Markers show Monte Carlo simulations, and solid lines show the analytical model [i.e., (16) and (18)]. Furthermore, dotted lines show the estimated upper bound [i.e., $\text{ABEP}_{u}^{U}$ in (22)]. The ABEP of user 1 (probe/intended link) is shown. It is worth mentioning that some simulation results (markers) are not shown due to the long simulation time for medium/high values of $E_m/N_0$. 
Fig. 16. ABEP of SSK (blue and green lines for $N_r = 1$ and $N_r = 3$, respectively) and QAM (red and magenta lines for $N_r = 1$ and $N_r = 3$, respectively) modulations with multiuser detection. Setup: $N_u = 2$; $\sigma_1^2 = 1$; and $\sigma_2^2 = 5 \times 10^{-2}$. Markers show Monte Carlo simulations, and solid lines show the analytical model (i.e., (16) and (18) for SSK modulation and the formula in the first row of Table II for QAM). The ABEP of user 1 (probe/intended link) is shown.
Fig. 17. ABEP of, on the left, (blue curves) SSK and (green curves) QAM, and, on the right, (red curves) PSK and (magenta curves) GSSK modulations with multiuser detection. Setup: $N_u = 2$; $\sigma^2 = 1$ and $\sigma^2 = 10^{-2}$; and $N_r = 3$. For GSSK modulation, we have $(N_t, N_{ta}) = (5, 2)$ if $N = 8$; $(N_t, N_{ta}) = (8, 4)$ if $N = 64$; and $(N_t, N_{ta}) = (11, 4)$ if $N = 256$. 
3-user scenario
The ABEP of each user is shown

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13. Relay-Aided SM
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The EARTH power model is a very simple and elegant model that relates the transmitted power of a BS to the total power consumed.


\[
P_{\text{supply}} = \begin{cases} 
N_{\text{RF}} P_0 + m N_{\text{RF}} P_t, & 0 < P_t \leq P_{\text{max}} \\
P_{\text{sleep}}, & P_t = 0. 
\end{cases}
\]

- \(P_{\text{supply}}\) is the total power supplied to the BS
- \(N_{\text{RF}}\) is the number of RF chains at the BS
- \(P_0\) is the power consumption per RF chain at the least transmission power
- \(m\) is the slope of the load-dependent power consumption
- \(P_t\) is the RF transmit-power per antenna
- \(P_{\text{max}}\) is the maximum transmit-power per antenna

### Power Model Parameters for Different BS (SOTA 2010)

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<tr>
<th>BS type</th>
<th>$P_0$ (W)</th>
<th>$m$</th>
<th>$P_{max}$ (W)</th>
<th>$P_{sleep}$ (W)</th>
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<td>40.0</td>
<td>63</td>
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<tr>
<td>Micro</td>
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<td>6.3</td>
<td>-</td>
</tr>
<tr>
<td>Pico</td>
<td>6.8</td>
<td>4.0</td>
<td>0.13</td>
<td>-</td>
</tr>
<tr>
<td>Femto</td>
<td>4.8</td>
<td>7.5</td>
<td>0.05</td>
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### Simulation Parameters

<table>
<thead>
<tr>
<th>BS Type</th>
<th>Values</th>
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<tbody>
<tr>
<td>Power Model Parameters</td>
<td>Macro, Micro, Pico, Femto</td>
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<td>Carrier Frequency</td>
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<td>Path Loss Model</td>
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<tr>
<td>Iterations (Number of Channels)</td>
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<tr>
<td>Bandwidth</td>
<td>10 MHz</td>
</tr>
<tr>
<td>Operating Temperature</td>
<td>Outdoor:290 K, Indoor: 293.5 K</td>
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</table>

### BS type | $d_{min}$ (m) | $d_{max}$ (m) |
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<tr>
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<td>1000</td>
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<td>Pico</td>
<td>10</td>
<td>150</td>
</tr>
<tr>
<td>Femto</td>
<td>5</td>
<td>50</td>
</tr>
</tbody>
</table>
\[ C_{SM} = C_1 + C_2 \approx C_1 = \frac{W}{N_t} \sum_{m=1}^{N_t} \log_2 \left( 1 + \frac{P}{N_0} \left\| h_m \right\|_2^2 \right) \]

\[ C_{STBC} = W R_{STBC} \log_2 \left( 1 + \frac{P}{N_0 N_t} \sum_{m=1}^{N_t} \left\| h_m \right\|_2^2 \right) \]

\[ C_{MISO \text{ (CSIT)}} = W \log_2 \left( 1 + \frac{P}{N_0} \left\| h \right\|_2^2 \right) \]

\[ \text{EE} = \frac{\text{Capacity}}{P_{\text{supply}}} \]

The following energy-model is considered:


\[ E_{tot} = \xi E_b \times \frac{(4\pi d)^2}{\eta G_t G_r \lambda^2} M_l N_f + \frac{P_{\text{circuit}}}{R_b} \]

- \( E_b \) is the bit energy
- \( R_b \) is the bit rate
- \( d \) is the transmission distance
- \( M_l \) is the link margin
- \( G_t \) and \( G_r \) are transmit and receive antenna gains
- \( N_f \) is the noise figure
- \( \lambda \) is the wavelength
- \( \eta \) is the drain efficiency of the power amplifier
- \( \xi \) is the peak-to-average-power-ratio (PAPR)
- \( P_{\text{circuit}} = P_{\text{DAC}} + P_{\text{mixer}} + P_{\text{filters}} + P_{\text{freqSynt}} \)

Energy Efficiency (9/26)

- The following energy-model is considered:
  - $E_b$ is the bit energy
  - $R_b$ is the bit rate
  - $d$ is the transmission distance
  - $M_l$ is the link margin
  - $G_t$ and $G_r$ are transmit and receive antenna gains
  - $N_f$ is the noise figure
  - $\lambda$ is the wavelength
  - $\eta$ is the drain efficiency of the power amplifier
  - $\xi$ is the peak-to-average-power-ratio (PAPR)
  - $P_{\text{circuit}} = P_{\text{DAC}} + P_{\text{mixer}} + P_{\text{filters}} + P_{\text{freqSynt}}$

<table>
<thead>
<tr>
<th>$f_c = 2.5$ GHz</th>
<th>$\eta = 0.35$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$d = 750$ m</td>
<td>Symbol Rate = 15 ksymbols/sec</td>
</tr>
<tr>
<td>$G_tG_r = 5$ dBi</td>
<td>$N_0 = -174$ dBm/Hz</td>
</tr>
<tr>
<td>$P_{\text{mix}} = 30.3$ mW</td>
<td>$M_l = 40$ dB</td>
</tr>
<tr>
<td>Target $\text{ABFP} = 10^{-5}$</td>
<td>$N_f = 10$ dB</td>
</tr>
<tr>
<td>$P_{\text{filt}} = 2.5$ mW</td>
<td>$P_{\text{syn}} = 50$ mW</td>
</tr>
</tbody>
</table>

SM vs. Single-RF QAM – 4 bpcu

Fig. 2. SM and single antenna/MRC comparison in Rayleigh channels in terms of: a) ABEP vs. $E_s/N_0$ and b) Relative energy gain vs. correlation factor.

SM vs. Single-RF QAM – 4 bpcu

SM vs. Single-RF QAM – 4 bpcu

Fig. 4. SM and single antenna/MRC comparison in Nakagami-$m$ channels with $m = 0.7$ in terms of: a) ABEP vs. $E_s/N_0$ and b) Relative energy gain vs. correlation factor.

SM vs. Single-RF QAM – 4 bpcu

Fig. 5. SM and single antenna/MRC comparison in Weibull channels with $b = 3$ in terms of: a) ABEP vs. $E_s/N_0$ and b) Relative energy gain vs. correlation factor.

Energy Efficiency (14/26)

SM vs. Single-RF QAM – 4 bpcu

![Graph showing energy efficiency comparison between SM and Single-RF QAM](image)

Fig. 6. SM and single antenna/MRC comparison in Weibull channels with $b = 1.5$ in terms of: a) ABEP vs. $E_s/N_0$ and b) Relative energy gain vs. correlation factor.

Energy efficiency is achieved by non-equiprobable signaling where less power-consuming modulation symbols are used more frequently to transmit a given amount of information.

The energy efficient modulation design is formulated as a convex optimization problem, where minimum achievable average symbol power consumption is derived with rate, performance, and hardware constraints.

Energy-Efficient Hamming Code-Aided (EE-HSSK) modulation

From GSSK …

Example of GSSK ($n_t = 2$) Modulation Alphabet and Bit Mapping for 3 Bits/s/Hz Transmission in a System with $N_T = 5$

<table>
<thead>
<tr>
<th>Source bits</th>
<th>GSSK symbols $\in A^{(GSSK)}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>000</td>
<td>$[0, 0, 0, 1, 1]^T$</td>
</tr>
<tr>
<td>001</td>
<td>$[0, 0, 1, 0, 1]^T$</td>
</tr>
<tr>
<td>010</td>
<td>$[0, 1, 0, 0, 1]^T$</td>
</tr>
<tr>
<td>011</td>
<td>$[1, 0, 0, 0, 1]^T$</td>
</tr>
<tr>
<td>100</td>
<td>$[0, 0, 1, 1, 0]^T$</td>
</tr>
<tr>
<td>101</td>
<td>$[0, 1, 0, 1, 0]^T$</td>
</tr>
<tr>
<td>110</td>
<td>$[1, 0, 0, 1, 0]^T$</td>
</tr>
<tr>
<td>111</td>
<td>$[0, 1, 1, 0, 0]^T$</td>
</tr>
</tbody>
</table>

- Limitations of GSSK:
  - Transmission rate
  - Selection of the spatial-constellation diagram
  - System performance ($d_{\text{min}} = 2$)

In HSSK:

- The set of antenna indices is fully utilized.
- It employs a different number of 1’s in each modulation symbol based on the Hamming code (in general, binary linear block code) construction technique.
- Increased number of RF chains.

---

Problem Formulation

- The objective of EE-HSSK modulation is to design an alphabet and the symbol a priori probabilities so that minimum average symbol power per transmission is achieved, while the target transmission rate (spectral-efficiency constraint), the minimum Hamming distance property (performance constraint), and the maximum required number of RF chains (hardware constraint) are met.

- Given a code $C = \{C_i\}$ with the specified minimum distance property.
- Given that each element in $C_i$ requires $i$ RF chains at the transmitter.
- Given that each element in $C_i$ consumes power equal to $i$.
- Given that the maximum number of RF chains is restricted to $i \leq M$.
- Then...

Problem Formulation

... the design problem is mathematically formulated as:

\[
\min_{P_i} \sum_{i:C_i \subseteq C, i \leq M} i |C_i| P_i \\
\text{s.t.} \sum_{i:C_i \subseteq C, i \leq M} |C_i| P_i = 1 \\
\sum_{i:C_i \subseteq C, i \leq M} |C_i| r(P_i) \geq m
\]

where \( P_i \) is the a priori probability of each symbol in \( C_i \), i.e., 
\( P_i \equiv P(x), x \in C_i \), and

\[
r(P_i) = \begin{cases} 
-P_i \log_2 P_i, & \text{if } P_i > 0 \\
0, & \text{if } P_i = 0 \\
-\infty, & \text{otherwise}
\end{cases}
\]

The a priori probabilities of all symbols in the alphabet sum to one, and \( P_i = 0 \) if \( i > M \)

The target information rate of \( m \) bits is met, as described by Shannon’s entropy formula

Optimal Solution

The optimization problem has a linear objective function subject to an affine equality and convex inequality constraints. Therefore, it is convex with a globally optimal solution, which can be found using the Lagrange multiplier method.

The optimal a priori transmission probabilities $P_i$ associated to the Lagrange multipliers $\lambda_1$ and $\lambda_2$ can be computed as follows:

$$i |C_i| + \lambda_1^* |C_i| + \lambda_2^* |C_i| \left( \log_2 P_i^* + \frac{1}{\log 2} \right) = 0. \quad (14)$$

Arranging the terms yields $P_i^* = \alpha \beta^i$, where $\alpha = (1/e)2^{-\lambda_1^*/\lambda_2^*}$ and $\beta = 2^{-1/\lambda_2^*}$. Normalizing $P_i^*$’s according to the equality constraint of Problem (9), we obtain

$$P_i^* = \frac{\beta^i}{\sum_{i : C_i \leq C, i \leq M} |C_i| \beta^i}, \quad 0 < \beta \leq 1. \quad (15)$$

Optimal Solution

- The value of $\beta$ determines the optimal a priori probabilities for the alphabet:
  - If $\beta = 1$, all codewords in C are included in the alphabet equiprobably to achieve the highest information rate. The cost is to have the largest average symbol power consumption.
  - If $\beta = 0^+$, only the least power-consuming codewords in C are included in the alphabet equiprobably.

- The solution provides the optimal symbol a priori probabilities. However, no information is given for accomplishing the bit mapping. Variable-length coding is proposed for creating an efficient bit-string representation of symbols with unequal a priori probabilities: Huffman coding.
  - The length of the bit strings is roughly reversely proportional to the symbol power. Since longer bit strings appear less frequently in a random input sequence, symbols more power-consuming are used less frequently to achieve energy efficiency.

Energy Efficiency (23/26)

\[ N_t = 7 \]
Energy Efficiency (24/26)

![Graph showing energy efficiency vs. transmission rate for different schemes]

\[ N_t = 10 \]
Energy Efficiency (25/26)

$N_t = N_r = 7$

![Graph showing SER versus $E_b/N_0$ for different modulation schemes.](image)

(Single RF-SIMO)
Energy Efficiency (26/26)

\[ N_t = N_r = 10 \]

(Single RF-SIMO)

(Two-RF-MIMO)
Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
8. Imperfect Channel State Information at the Receiver
9. Multiple Access Interference
10. Energy Efficiency
11. Transmit-Diversity for SM
12. Spatially-Modulated Space-Time-Coded MIMO
13. Relay-Aided SM
14. SM in Heterogeneous Cellular Networks
15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
The Alamouti Scheme

\[ g_2 = \begin{pmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{pmatrix} \]

Orthogonal Space-Time Block Codes (OSTBCs)

\[
\mathcal{H}_3 = \begin{bmatrix}
\frac{x_3}{\sqrt{2}} & \frac{-x_2^*}{\sqrt{2}} & \frac{x_1}{\sqrt{2}} & \frac{x_2}{\sqrt{2}} & \frac{x_3}{\sqrt{2}} \\
\frac{x_1^*}{x_1} & \frac{x_3^*}{x_3} & \frac{x_2^*}{x_2} & \frac{x_3^*}{x_3} & \frac{x_2^*}{x_2} \\
\frac{x_3^*}{\sqrt{2}} & \frac{x_2^*}{\sqrt{2}} & \frac{-x_1^*}{\sqrt{2}} & \frac{-x_2^*}{\sqrt{2}} & \frac{-x_3^*}{\sqrt{2}} \\
\end{bmatrix}
\]

\[
\mathcal{H}_4 = \begin{bmatrix}
\frac{x_3}{\sqrt{2}} & \frac{-x_2^*}{\sqrt{2}} & \frac{x_1}{\sqrt{2}} & \frac{x_2}{\sqrt{2}} & \frac{x_3}{\sqrt{2}} \\
\frac{x_1^*}{x_1} & \frac{x_3^*}{x_3} & \frac{x_2^*}{x_2} & \frac{x_3^*}{x_3} & \frac{x_2^*}{x_2} \\
\frac{x_3^*}{\sqrt{2}} & \frac{x_2^*}{\sqrt{2}} & \frac{-x_1^*}{\sqrt{2}} & \frac{-x_2^*}{\sqrt{2}} & \frac{-x_3^*}{\sqrt{2}} \\
\frac{x_2^*}{\sqrt{2}} & \frac{x_3^*}{x_3} & \frac{x_1^*}{x_1} & \frac{x_2^*}{x_2} & \frac{x_3^*}{x_3} \\
\frac{x_3^*}{\sqrt{2}} & \frac{x_2^*}{\sqrt{2}} & \frac{-x_1^*}{\sqrt{2}} & \frac{-x_2^*}{\sqrt{2}} & \frac{-x_3^*}{\sqrt{2}} \\
\end{bmatrix}
\]

Opportunity: Transmit-diversity with rate greater than one

Challenge: Transmit-diversity with rate greater than one and single-stream decoding complexity
Transmit-Diversity for SM (4/61)

\[
\begin{align*}
 h_1(t) &= \beta_1 \exp(j\varphi_1) \delta(t - \tau_0) \\
 h_2(t) &= \beta_2 \exp(j\varphi_2) \delta(t - \tau_0) \\
 \tilde{s}_1(t | \{m_n\}_{n=1}^2) &= \sqrt{E_m} \beta_1 \exp(j\varphi_1) s_1(t | \{m_n\}_{n=1}^2) \\
 \tilde{s}_2(t | \{m_n\}_{n=1}^2) &= \sqrt{E_m} \beta_2 \exp(j\varphi_2) s_2(t | \{m_n\}_{n=1}^2) \\
 r(t | m_1) &= \tilde{s}_1(t | m_1) + \tilde{s}_2(t | m_1) + n(t) = \overline{s}_1(t) + n(t) \\
 r(t | m_2) &= \tilde{s}_1(t | m_2) + \tilde{s}_2(t | m_2) + n(t) = \overline{s}_2(t) + n(t) \\
 \hat{m} &= \begin{cases} 
 m_1 & \text{if } D_1 \geq D_2 \\
 m_2 & \text{if } D_2 < D_1 
\end{cases} \\
 D_1 &= \text{Re}\left\{ \int_{T_m} r(t) \overline{s}_1^*(t) \, dt \right\} - \frac{1}{2} \int_{T_m} \overline{s}_1(t) \overline{s}_1^*(t) \, dt \\
 D_2 &= \text{Re}\left\{ \int_{T_m} r(t) \overline{s}_2^*(t) \, dt \right\} - \frac{1}{2} \int_{T_m} \overline{s}_2(t) \overline{s}_2^*(t) \, dt
\end{align*}
\]

Transmit-Diversity for SM (5/61)

- Transmitted Signal:
  - If $m_1$ needs to be transmitted: TX$_1$ is active and TX$_2$ radiates no power
  - If $m_2$ needs to be transmitted: TX$_1$ and TX$_2$ are both active

  $$
  \begin{align*}
  s_1(t|m_1) &= s_1(t|m_2) = s_2(t|m_2) = 1 \\
  s_2(t|m_1) &= 0
  \end{align*}
  $$

- Received Signal:

  $$
  \begin{align*}
  r(t|m_1) &= \sqrt{E_m}\beta_1 \exp(j\varphi_1) + n(t) \\
  r(t|m_2) &= \sqrt{E_m}\beta_1 \exp(j\varphi_1) + \sqrt{E_m}\beta_2 \exp(j\varphi_2) + n(t)
  \end{align*}
  $$

- Error Probability:

  $$
  BEP = Q\left(\sqrt{\frac{E_m}{4N_0} \beta_2^2}\right) \quad \Rightarrow \quad ABEP = \frac{1}{2} - \frac{1}{2} \sqrt{\frac{\sigma_2^2(E_m/4N_0)}{1 + \sigma_2^2(E_m/4N_0)}}
  $$

Transmit-Diversity for SM (7/61)

- **Transmitted Signal:**
  - If $m_1$ needs to be transmitted: TX$_1$ is active and TX$_2$ radiates no power
  - If $m_2$ needs to be transmitted: TX$_1$ radiates no power and TX$_2$ is active

\[
\begin{align*}
    s_1(t|m_1) &= s_2(t|m_2) = 1 \\
    s_1(t|m_2) &= s_2(t|m_1) = 0
\end{align*}
\]

- **Received Signal:**

\[
\begin{align*}
    r(t|m_1) &= \sqrt{E_m} \beta_1 \exp(j\varphi_1) + n(t) \\
    r(t|m_2) &= \sqrt{E_m} \beta_2 \exp(j\varphi_2) + n(t)
\end{align*}
\]

- **Error Probability:**

\[
\text{BEP} = Q\left(\sqrt{\frac{E_m}{4N_0}} |\beta_2 \exp(j\varphi_2) - \beta_1 \exp(j\varphi_1)|^2\right) \quad \Rightarrow \quad \text{ABEP} = \frac{1}{2} - \frac{1}{2} \sqrt{\frac{\tilde{\sigma}^2 (E_m/4N_0)}{1 + \tilde{\sigma}^2 (E_m/4N_0)}}
\]

Transmit-Diversity for SM (9/61)

- **Transmitted Signal (TOSD-SSK):**
  - If \( m_1 \) needs to be transmitted: TX\(_1\) is active and TX\(_2\) radiates no power
  - If \( m_2 \) needs to be transmitted: TX\(_1\) radiates no power and TX\(_2\) is active

\[
\begin{align*}
  s_1(t | m_1) &= w_1(t) \\
  s_2(t | m_2) &= w_2(t) \\
  s_1(t | m_2) &= s_2(t | m_1) = 0
\end{align*}
\]

- **Received Signal:**

\[
\begin{align*}
  r(t | m_1) &= \sqrt{E_m} \beta_1 \exp(j \varphi_1) w_1(t) + n(t) \\
  r(t | m_2) &= \sqrt{E_m} \beta_2 \exp(j \varphi_2) w_2(t) + n(t)
\end{align*}
\]

- **Error Probability:**

\[
\text{BEP} = Q\left(\sqrt{\frac{E_m}{4N_0}} (\beta_1^2 + \beta_2^2)\right) \quad \Rightarrow \quad \begin{cases} 
  \text{ABEP} = \frac{1}{\pi} \int_{0}^{\pi/2} M \left( \frac{E_m / 4N_0}{2 \sin^2(\theta)} \right) d\theta \\
  M(s) = \left[ 1 + 2(\sigma_1^2 + \sigma_2^2) s + 4(1 - \rho^2) \sigma_1^2 \sigma_2^2 s^2 \right]^{-1}
\end{cases}
\]
Transmit-Diversity for SM (10/61)

- If $w_1(t) = w_2(t) \rightarrow \text{Diversity} = 1$ (conventional SSK)
- If $w_1(t)$ is “*time-orthogonal*” to $w_2(t) \rightarrow \text{Diversity} = 2$ (TOSD-SSK)

[1] Chau and Yu

[3]-[5]: Mesleh et al. and Jeganathan et al.

TOSD-SM: Time-Orthogonal Signal Design assisted SM

Generalization to Rician Fading, $N_t > 2$, and $N_r > 1$

- If $w_i(t) = w_j(t)$, Diversity = $N_r$ (conventional SSK)
- If $w_i(t)$ is “time-orthogonal” to $w_j(t)$, Diversity = $2N_r$ (TOSD-SSK)
- This is true for any $N_t$ with no bandwidth expansion and with a single active transmit-antenna at any time-instance

Orthogonal Waveforms Design with Bandwidth Constraint

\[
\begin{align*}
p_1(\xi) &= \left(\frac{1}{\sqrt{\sqrt{\pi}}}\right) \exp \left[-\frac{1}{2} \left(\frac{\xi}{t_0}\right)^2\right] \\
p_2(\xi) &= \left(\frac{2\xi}{\sqrt{2\sqrt{\pi}}}\right) \exp \left[-\frac{1}{2} \left(\frac{\xi}{t_0}\right)^2\right] \\
p_3(\xi) &= \left(\frac{4\xi^2-2}{\sqrt{8\sqrt{\pi}}}\right) \exp \left[-\frac{1}{2} \left(\frac{\xi}{t_0}\right)^2\right] \\
p_4(\xi) &= \left(\frac{8\xi^3-12\xi}{\sqrt{18\sqrt{\pi}}}\right) \exp \left[-\frac{1}{2} \left(\frac{\xi}{t_0}\right)^2\right] \\
p_5(\xi) &= \left(\frac{16\xi^4-48\xi^2+12}{\sqrt{384\sqrt{\pi}}}\right) \exp \left[-\frac{1}{2} \left(\frac{\xi}{t_0}\right)^2\right]
\end{align*}
\]

\[
\begin{align*}
w_{t1}(\xi) &= \left(-\frac{4}{\sqrt{165}}\right) p_1(\xi) + \left(\frac{\sqrt{4-2\sqrt{2}}}{4}\right) p_2(\xi) + \left(\frac{\sqrt{11}}{30}\right) p_3(\xi) + \left(-\frac{\sqrt{4+2\sqrt{2}}}{4}\right) p_4(\xi) + \left(-\frac{2}{\sqrt{110}}\right) p_5(\xi) \\
w_{t2}(\xi) &= \left(-\frac{4}{\sqrt{165}}\right) p_1(\xi) + \left(-\frac{\sqrt{4-2\sqrt{2}}}{4}\right) p_2(\xi) + \left(\frac{\sqrt{11}}{30}\right) p_3(\xi) + \left(\frac{\sqrt{4+2\sqrt{2}}}{4}\right) p_4(\xi) + \left(-\frac{2}{\sqrt{110}}\right) p_5(\xi) \\
w_{t3}(\xi) &= \left(\frac{3}{22}\right) p_1(\xi) + \left(-\frac{\sqrt{4+2\sqrt{2}}}{4}\right) p_2(\xi) + (0) p_3(\xi) + \left(\frac{\sqrt{4-2\sqrt{2}}}{4}\right) p_4(\xi) + \left(-\frac{2}{\sqrt{11}}\right) p_5(\xi) \\
w_{t4}(\xi) &= \left(\frac{3}{22}\right) p_1(\xi) + \left(\frac{\sqrt{4+2\sqrt{2}}}{4}\right) p_2(\xi) + (0) p_3(\xi) + \left(-\frac{\sqrt{4-2\sqrt{2}}}{4}\right) p_4(\xi) + \left(-\frac{2}{\sqrt{11}}\right) p_5(\xi)
\end{align*}
\]


Transmit-Diversity for SM (15/61)

\[
FPCB_{X \%} = \min_{B \in [0, +\infty)} \left\{ B \left( \int_{0}^{B} \frac{|P(\omega)|^2 d\omega}{\int_{0}^{\infty} |P(\omega)|^2 d\omega} \right) > X \% \right\}
\]

\[
BPSDB_{TH_{dB}} = \min_{B \in [0, +\infty)} \left\{ B \log_{10} \left( |P(\omega)|^2 \right) < \log_{10} \left( |P(\omega_{\text{peak}})|^2 \right) - TH_{dB}, \forall \omega > B \right\}
\]

Transmit-Diversity for SM (16/61)

Fig. 5. TOSD–SSK modulation: ABEP against $E_m/N_0$. Solid, dashed, and dotted lines denote the analytical model in Section IV and markers Monte Carlo simulations. Setup: i) $N_t = 2$, ii) $N_r = 2$, iii) $\Omega_{i,t} = 10\text{dB}$ and $K_{R}^{(i,l)} = K_R$ for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$, and iv) $\rho = 0.00$ (solid lines), $\rho = 0.25$ (dashed lines), $\rho = 0.75$ (dotted lines).
Fig. 7. TOSD–SSK modulation: ABEP against $E_m/N_0$. Solid, dashed, and dotted lines denote the analytical model in Section IV and markers Monte Carlo simulations. Setup: i) $N_t = 4$, ii) $N_r = 2$, iii) $\Omega_{i,l} = 10$ dB and $K^{(i,l)}_R = K_R$ for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$, and iv) $\rho = 0.00$ (solid lines), $\rho = 0.25$ (dashed lines), $\rho = 0.75$ (dotted lines).
Fig. 8. TOSD-SSK modulation: ABEP against $E_m/N_0$. Solid, dashed, and dotted lines denote the analytical model in Section IV and markers Monte Carlo simulations. Setup: i) $N_t = 4$, ii) $N_r = 4$, iii) $\Omega_{i,t} = 10$dB and $K_{R(i,l)} = K_R$ for $i = 1, 2, \ldots, N_t$, and $l = 1, 2, \ldots, N_r$, and iv) $\rho = 0.00$ (solid lines), $\rho = 0.25$ (dashed lines), $\rho = 0.75$ (dotted lines).
Fig. 11. Comparison between SSK and TOSD-SSK modulation: ABEP against $E_m/N_0$. Markers with solid lines denote the analytical model in Section III and markers with dotted lines the analytical model in Section IV. Setup: i) $N_t = 2$, ii) $\Omega_{i,l} = 10$dB and $K_R^{(i,l)} = 5$dB for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$, and iii) $\rho = 0.25$. 

Transmit-Diversity for SM (19/61)
Fig. 12. Comparison between SSK and TOSD-SSK modulation: ABEP against $E_m/N_0$. Markers with solid lines denote the analytical model in Section III and markers with dotted lines the analytical model in Section IV. Setup: i) $N_r = 2$, and ii) $\Omega_{i,l} = 10$dB and $K_{R}^{(i,l)} = 0$dB for $i = 1, 2, \ldots, N_t$ and $l = 1, 2, \ldots, N_r$. 
In summary:

- TOSD-SSK achieves transmit-diversity with just 1 active antenna at the transmitter
- However, TOSD-SSK achieves transmit-diversity only equal to 2 \( \rightarrow \) Full transmit-diversity is possible only if \( N_t = 2 \)
- Furthermore, the data rate of SSK is only \( \text{Rate}=\log_2(N_t) \) \( \rightarrow \) This is too low for high data rate applications

Questions:

- Can we achieve a transmit-diversity gain greater than 2?
- At the same time, can we increase the rate?
- Given a pair (rate, diversity), how to design a SSK scheme achieving it?
Increasing the Rate via GSSK

- Size of the spatial-constellation diagram ($N_H > N_t$)

$$N_H = 2^{\left\lceil \log_2 \left( \frac{N_t}{N_a} \right) \right\rceil}$$

- Rate $= \log_2 (N_H) > \log_2 (N_t)$

- Spatial-constellation diagram:
  - $N_a = 1$ (i.e., SSK) $\rightarrow$ $D = \{1, 2, 3, 4, 5\}$
  - $N_a = 2$ $\rightarrow$ $D = \{(1,2); (1,3); (1,4); (1,5); (2,3); (2,4); \ldots\}$
  - $N_a = 3$ $\rightarrow$ $D = \{(1,2,3); (1,2,4); (1,2,5); (1,3,4); \ldots\}$

Problem statement

- Let $N_t$ be the transmit-antennas and $N_a$ be the active transmit-antennas.
- Then, the largest possible size of the spatial-constellation diagram is:

$$\tilde{N}_H = 2^{\left\lfloor \log_2 \left( \frac{N_t}{N_a} \right) \right\rfloor}$$

Objectives

- Find the actual spatial constellation diagram of size $N_H \leq \tilde{N}_H$ such that transmit-diversity is Div.
- Understand the role played by the TOSD principle for transmit-diversity.

Methodology

- We have computed the PEP (Pairwise Error Probability) of any pair of points in the spatial-constellation diagram and have analyzed the transmit-diversity order of each of them.

Transmit-Diversity for SM (24/61)

Main Result: Transmit-Diversity 1 and 2

- **Result 1 (Div=1)**
  - The system achieves transmit-diversity $\text{Div}=1$ and rate $R=\log_2(N_H)$ if the $N_t$ transmit-antennas have the same shaping filter
  - This scheme is called GSSK and reduces to SSK if $N_a=1$

- **Result 2 (Div=2)**
  - The system achieves transmit-diversity $\text{Div}=2$ and rate $R=\log_2(N_H)$ if the $N_t$ transmit-antennas have orthogonal shaping filters
  - This scheme is called TOSD-GSSK and reduces to TOSD-SSK if $N_a=1$

Main Result: Transmit-Diversity $> 2$

- **Result 3 ($Div > 2$)**
  - Let $N_H^\perp$ be the size of the partition of the set of $N_t$ transmit-antennas such that $N_t = N_H^\perp \cdot N_a \Rightarrow$ each subset of the partition has $N_a$ distinct antenna-elements and the subsets are pairwise disjoint.
  - Then, the system achieves transmit-diversity $Div = 2 \cdot N_a$ and rate $R = \log_2(N_H^\perp)$ if the $N_t$ transmit-antennas have orthogonal shaping filters.
  - This scheme is called TOSD-GSSK with mapping by pairwise disjoint set partitioning (TOSD-GSSK-SP).

\[
R = \log_2 \left( \frac{N_t}{N_a} \right) \quad \text{tradeoff} \quad \iff \quad Div = 2N_a
\]

Transmit-Diversity for SM (26/61)

$N_t=4$, $N_a=2$, $R=1$, $Div=4$

Five schemes are studied:

- **SSK**: $N_a = 1$, $w_0(.) = w_i(.)$, $\text{Div}=1$

- **GSSK**: $N_a > 1$, $w_0(.) = w_i(.)$, $\text{Div}=1$

- **TOSD-SSK**: $N_a = 1$, $N_t$ orthogonal $w_i(.)$, $\text{Div}=2$

- **TOSD-GSSK**: $N_a > 1$, $N_t$ orthogonal $w_i(.)$, $\text{Div}=2$

- **TOSD-GSSK-SP**: $N_a > 1$, $N_t$ orthogonal $w_i(.)$, the spatial-constellation diagram is a partition of $N_t$, $\text{Div}=2 \cdot N_a$

Div = 1 and Div = 2

$10^{-1}$

$10^{-2}$

$10^{-3}$

$10^{-4}$

$10^{-5}$

$E_m/N_0$ [dB]

GSSK [Nt=5, Na=2, R=3]

GSSK [Nt=6, Na=3, R=4]

TOSD-GSSK [Nt=5, Na=2, R=3]

TOSD-GSSK [Nt=6, Na=3, R=4]
Transmit-Diversity for SM (29/61)

\[ R = 1 - \text{TOSD-GSSK-SP} \]

\[ N_t = 4, \quad N_a = 2, \quad \text{Div} = 4 \]

\[ N_t = 6, \quad N_a = 3, \quad \text{Div} = 6 \]

\[ N_t = 8, \quad N_a = 4, \quad \text{Div} = 8 \]
Transmit-Diversity for SM (30/61)

R = 1

$\begin{align*}
\text{SSK} \ [N_t=2, \ Na=1, \ Div=1] \\
\text{TOSD-SSK} \ [N_t=2, \ Na=1, \ Div=2] \\
\text{TOSD-GSSK-SP} \ [N_t=4, \ Na=2, \ Div=4] \\
\text{TOSD-GSSK-SP} \ [N_t=6, \ Na=3, \ Div=6] \\
\text{TOSD-GSSK-SP} \ [N_t=8, \ Na=4, \ Div=8]
\end{align*}$

$E_m/N_0$ [dB]

$\text{ABEP}$
Transmit-Diversity for SM (31/61)

\[ R = 2 \]

- SSK \([N_t=4, Na=1, Div=1]\)
- TOSD-SSK \([N_t=4, Na=1, Div=2]\)
- TOSD-GSSK-SP \([N_t=8, Na=2, Div=4]\)
- TOSD-GSSK-SP \([N_t=12, Na=3, Div=6]\)

Graph showing the relationship between \( E_m / N_0 \) in dB and ABEP.
Transmit-Diversity for SM (32/61)

Nt = 8

- SSK [Na=1, R=3, Div=1]
- TOSD-SSK [Na=1, R=3, Div=2]
- GSSK [Na=4, R=6, Div=1]
- TOSD-GSSK [Na=4, R=6, Div=2]
- TOSD-GSSK-SP [Na=2, R=2, Div=4]
- TOSD-GSSK-SP [Na=4, R=1, Div=8]

Em/N₀ [dB] vs. ABEP

Graph showing performance comparison of various transmit diversity techniques.
Transmit-Diversity for SM (33/61)

\[ \text{Na} = 3 \]

\[
\begin{align*}
\text{GSSK} & \quad [N_t=6, \ R=4, \ \text{Div}=1] \\
\text{GSSK} & \quad [N_t=7, \ R=5, \ \text{Div}=1] \\
\text{TOSD-GSSK} & \quad [N_t=6, \ R=4, \ \text{Div}=2] \\
\text{TOSD-GSSK} & \quad [N_t=7, \ R=5, \ \text{Div}=2] \\
\text{TOSD-GSSK-SP} & \quad [N_t=6, \ R=1, \ \text{Div}=6] \\
\text{TOSD-GSSK-SP} & \quad [N_t=12, \ R=2, \ \text{Div}=6]
\end{align*}
\]
Transmit-Diversity for SM (34/61)

- From SSK to SM
  - Understanding the design challenges of transmit-diversity for SM
  - Generalizing the TOSD approach to SM (TOSD-SM)
  - Interested in transmit-diversity equal to 2 (extension of Alamouti code)

- Challenges (…let us start, e.g., from Alamouti…)

Problem statement

- Let $N_t$ be the transmit-antennas and $N_a$ be the active transmit-antennas
- Then, the largest possible size of the spatial-constellation diagram is:

$$N_H = 2 \log_2 \left( \frac{N_t}{N_a} \right)$$

**Objective.** Find the actual spatial constellation diagram of size $N_h \leq N_H$ such that:

- Transmit-diversity is 2 for $N_a = 2$
- Transmit-diversity can be achieved with single-stream decoding complexity

Methodology

- We have computed the PEP (Pairwise Error Probability) of any pair of (antenna-index, modulated-symbol) and have analyzed transmit-diversity and single-stream decoding optimality of each of them

Main Result: Same Shaping Filters at Tx

- **Result 1 (receiver complexity)**
  - Whatever the spatial-constellation diagram is, if the shaping filters at the transmitter are all the same, adding the SSK component on top of the Alamouti code destroys its inherent orthogonality. So, no single-stream decoder can be used and the receiver complexity is of the order of \( N_h \cdot M^N_a \) correlations.

- **Result 2 (transmit-diversity)**
  - If the shaping filters at the transmitter are all the same, transmit-diversity equal to 2 can be guaranteed by partitioning the spatial-constellation diagram into non-overlapping sets of antennas. However, a multi-stream receiver is needed at the destination for ML-optimum decoding.

From Result 1 and Result 2, it follows that this scheme achieves transmit-diversity equal to 2 but multi-stream decoding is needed.

Main Result: Time-Orthogonal Shaping Filters at Tx

- **Result 3 (receiver complexity)**
  - ML-optimum low-complexity single-stream decoding can be guaranteed via an adequate choice of the (precoding) shaping filters at the transmitter. In particular, some pairs of filters should have zero cross-correlation function.

- **Result 4 (transmit-diversity)**
  - ML-optimum low-complexity single-stream decoding with transmit-diversity of 2 can be guaranteed via an adequate choice of both the precoding shaping filters and the spatial-constellation diagram at the transmitter. In particular, some pairs of filters must have zero cross-correlation function, and the spatial-constellation diagram should be a partition of the transmit-antenna array.

From Result 3 and Result 4, it follows that this scheme achieves transmit-diversity equal to 2 with single-stream decoding.

Transmit-Diversity for SM (40/61)

- **Case studies**
  - **Worst-case setup**, which achieves transmit-diversity equal to 1 and needs a multi-stream decoder at the destination. It is obtained by using the same shaping filters in all the antennas at the transmitter along with a spatial-constellation diagram with overlapping sets of points (SM-STBC).
  - **Best-case setup**, which achieves transmit-diversity equal to 2 and needs a single-stream decoder at the destination. This is obtained by using different and time-orthogonal shaping filters at the transmitter along with a spatial-constellation diagram composed by non-overlapping sets of points (TOSD-SM-STBC).

- **Baseline schemes**
  - SM
  - Alamouti code (rate=1)
  - H3 and H4 OSTBCs (rate=3/4)

Transmit-Diversity for SM (41/61)

![Graph showing the performance of different transmit diversity schemes for SM systems. The x-axis represents the signal-to-noise ratio (Em/N0) in dB, and the y-axis represents the average bit error probability (ABEP). Various schemes are compared, including Alamouti [M=8], SM [Nt=2, M=4], SM [Nt=4, M=2], SM-STBC [Nt=4, Nh=4, M=4], SM-STBC [Nt=7, Nh=16, M=2], and TOSD-SM-STBC [Nt=8, Nh=4, M=4]. The graph illustrates the trade-offs and performance gains for different modulation schemes and transmit diversity techniques.]
Transmit-Diversity for SM (42/61)

- 10^0
- 10^-1
- 10^-2
- 10^-3
- 10^-4
- 10^-5

- 0 5 10 15 20 25 30 35 40 45 50

- ABEP
- E_m/N_0 [dB]

- Alamouti [M=32]
- SM [Nt=2, M=16]
- SM [Nt=8, M=4]
- SM-STBC [Nt=4, Nh=4, M=16]
- SM-STBC [Nt=7, Nh=16, M=8]
- TOSD-SM-STBC [Nt=8, Nh=4, M=16]
Transmit-Diversity for SM (43/61)

1.5 bits/s/Hz

- STBC-H3 [M=4]
- STBC-H4 [M=4]
- SM-STBC [Nt=3, Nh=2, M=2]
- TOSD-SM-STBC [Nt=4, Nh=2, M=2]

Plot shows the relationship between \( E_m/N_0 [\text{dB}] \) and ABEP.
Transmit-Diversity for SM (44/61)

4.5 bits/s/Hz

EM/N0 [dB]

STBC-H3 [M=64]
STBC-H4 [M=64]
SM-STBC [Nt=3, Nh=2, M=16]
SM-STBC [Nt=5, Nh=8, M=8]
TOSD-SM-STBC [Nt=4, Nh=2, M=16]
Transmit-Diversity for SM (46/61)

Example:
- $N_t = 4$
- BPSK Alamouti
- $R = 2$ bpcu

\[
\chi_1 = \begin{pmatrix} x_1 & x_2 & 0 & 0 \\ -x_2^* & x_1^* & 0 & 0 \end{pmatrix}, \begin{pmatrix} 0 & 0 & x_1 & x_2 \\ 0 & 0 & -x_2^* & x_1^* \end{pmatrix}
\]

\[
\chi_2 = \begin{pmatrix} 0 & x_1 & x_2 & 0 \\ 0 & -x_2^* & x_1^* & 0 \end{pmatrix}, \begin{pmatrix} x_2 & 0 & 0 & x_1 \\ x_1^* & 0 & 0 & -x_2^* \end{pmatrix}
\]

$e^{j\theta}$

Fig. 3. Block diagram of the STBC-SM ML receiver.

Fig. 5. BER performance at 3 bits/s/Hz for STBC-SM, SM, V-BLAST, OSTBC and Alamouti’s STBC schemes.
Fig. 6. BER performance at 4 bits/s/Hz for STBC-SM, SM, V-BLAST, OSTBC and Alamouti’s STBC schemes.
Fig. 7. BER performance at 5 bits/s/Hz for STBC-SM, SM, V-BLAST, OSTBC and Alamouti’s STBC schemes.
Transmit-Diversity for SM (51/61)

Fig. 8. BER performance at 6 bits/s/Hz for STBC-SM, SM, V-BLAST, OSTBC and Alamouti’s STBC schemes.
The codewords $X$ of the Golden Code are 2x2 complex matrices of the following form:

$$X = \frac{1}{\sqrt{5}} \begin{bmatrix} \alpha [a+b\theta] & \alpha [c+d\theta] \\ i \sigma(\alpha) [c+d\sigma(\theta)] & \sigma(\alpha) [a+b\sigma(\theta)] \end{bmatrix}$$

where

- $a, b, c, d$ are the information symbols which can be taken from any M-QAM constellation carved from $\mathbb{Z}[i]$
- $i = \sqrt{-1}$
- $\theta = (1+\sqrt{5})/2 = 1.618...$ (Golden number)
- $\sigma(\theta) = (1-\sqrt{5})/2 = 1-\theta$
- $\alpha = 1 + i \cdot i \theta = 1 + i \sigma(\theta)$
- $\sigma(\alpha) = 1 + i \cdot i \sigma(\theta) = 1 + i \theta$


Double Space-Time Transmit Diversity (DSTTD)

Fig. 9. BER performance for STBC-SM, the Golden code and DSTTD schemes at 4 and 6 bits/s/Hz spectral efficiencies.
Transmit-Diversity for SM (55/61)

Fig. 10. BER performance at 3 bits/s/Hz for STBC-SM, SM, and Alamouti's STBC schemes for SC channel with $r = 0, 0.5$ and $0.9$. 
SM-CIOD: Transmit-Diversity with a Single-RF Chain

\[
\begin{bmatrix}
\tilde{s}_1 & 0 \\
0 & \tilde{s}_2
\end{bmatrix}
\]

channel uses

\[
s = \exp\left( j \frac{\arctan(2)}{2} \right)x_{QAM}
\]

\[
\begin{aligned}
\tilde{s}_1 &= s_{1,I} + js_{2,Q} \\
\tilde{s}_2 &= s_{2,I} + js_{1,Q}
\end{aligned}
\]

SM-CIOD: Transmit-Diversity with a Single-RF Chain

- First channel use: antenna 1 is used
- Second channel use: antenna \((l+1) \mod N_t\) is used

SM-CIOD: Transmit-Diversity with a Single-RF Chain

\[ CBS_i = \{ e^{j \theta_i} CB_{i,j} \mid 1 \leq j \leq N_t \} \text{ for } 1 \leq i \leq N_t, \]

<table>
<thead>
<tr>
<th>Modulation</th>
<th>$e^{j\theta_1}$</th>
<th>$e^{j\theta_2}$</th>
<th>$e^{j\theta_3}$</th>
<th>$e^{j\theta_4}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>4-QAM</td>
<td>$-0.9239 + 0.3827j$</td>
<td>$1.0000j$</td>
<td>$0.7071 + 0.7071j$</td>
<td>$0.9239 + 0.3827j$</td>
</tr>
<tr>
<td>8-QAM</td>
<td>$1$</td>
<td>$0.8090 - 0.5878j$</td>
<td>$-0.3090 + 0.951j$</td>
<td>$0.3090 + 0.951j$</td>
</tr>
</tbody>
</table>

Transmit-Diversity for SM (60/61)

Low DoSM scheme with $N_t=4$, $N_r=2$, 8-QAM ($R=4$ bpcu)

High DoSM scheme with $N_t=5$, $N_r=2$, 4-QAM ($R=4$ bpcu)
Transmit-Diversity for SM (61/61)

Low DoSM scheme with $N_t=4$, $N_r=2$, 16-QAM (R= 5 bpcu)

High DoSM scheme with $N_t=5$, $N_r=2$, 8-QAM (R= 5 bpcu)
Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
8. Imperfect Channel State Information at the Receiver
9. Multiple Access Interference
10. Energy Efficiency
11. Transmit-Diversity for SM
12. Spatially-Modulated Space-Time-Coded MIMO
13. Relay-Aided SM
14. SM in Heterogeneous Cellular Networks
15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
Opportunities and Challenges for SM

- **Opportunity**: Transmit-diversity with rate greater than one
- **Challenge**: Transmit-diversity with rate greater than one and single-stream decoding complexity
The signal received at the $r$–th receive–antenna and at the $s$–th time–slot is $((s - 1) T_s \leq \xi < s T_s)$:

$$z_{s,r} (\xi) = \sqrt{\frac{E_S}{\|a^{(\alpha)}\|_F^2}} \sum_{t=1}^{N_t} \left[ X_{s,t}^{(\alpha)} (\mu) H_{r,t} w_t (\xi) \right] + n_{s,r} (\xi)$$

where we have defined:

$$X_{s,t}^{(\alpha)} (\mu) = a_t^{(\alpha)} M_{s,t \circ N_\alpha} (\mu)$$

$$= \begin{cases} 0 & \text{if } a_t^{(\alpha)} = 0 \\ M_{s,t \circ N_\alpha} (\mu) & \text{if } a_t^{(\alpha)} = 1 \end{cases}$$

- $N_t$ transmit-antennas
- $N_\alpha$ active transmit-antennas
- $N_r$ receive-antennas
- $N_s$ time-slots

Spatially-Modulated Space-Time-Coded MIMO (4/23)

\[
(\hat{\alpha}, \hat{\mu}) = \arg\min_{a(\hat{\alpha}) \in A, \hat{\mu} \in \{\hat{\mu}_1, \ldots, \hat{\mu}_{NM}\} \in M} \left\{ \sum_{s=1}^{N_s} \sum_{r=1}^{N_r} \int_{(s-1)T_s}^{sT_s} z_{s,r}(\xi) - \sqrt{\frac{E_S}{\|a(\hat{\alpha})\|^2_F}} \sum_{l=1}^{N_t} \left| X_{s,r}^{(\hat{\alpha})}(\hat{\mu}) H_{r,t} w_t(\xi - (s-1)T_s) \right|^2 d\xi \right\}
\]

\[
\text{ABEP} \leq \sum_{\chi} \frac{1}{AM^{NM}} \sum_{\tilde{\chi} \neq \chi} \frac{H_d(\chi, \tilde{\chi}) \text{APEP}(\chi \rightarrow \tilde{\chi})}{\log_2(A) + NM \log_2(M)}
\]

\[
\gamma_{\chi \rightarrow \tilde{\chi}}(H) = \sum_{s=1}^{N_s} \sum_{r=1}^{N_r} \left\{ \sum_{t_1=1}^{N_t} \sum_{t_2=1}^{N_t} \left[ \frac{X_{s,t_1}^{(\hat{\alpha})}(\hat{\mu})}{\|a(\hat{\alpha})\|^2_F} - \frac{X_{s,t_1}^{(\alpha)}(\mu)}{\|a(\alpha)\|^2_F} \right] \left( \frac{X_{s,t_2}^{(\hat{\alpha})}(\hat{\mu})}{\|a(\hat{\alpha})\|^2_F} - \frac{X_{s,t_2}^{(\alpha)}(\mu)}{\|a(\alpha)\|^2_F} \right) \right)^* H_{r,t_1} H_{r,t_2}^* \int_0^{T_s} w_{t_1}(\xi) w_{t_2}(\xi) d\xi \right\}
\]

\[
\gamma_{\chi \rightarrow \tilde{\chi}}(H) = \eta(H)^H \Psi(\chi, \tilde{\chi}) \eta(H)
\]


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\[ \text{APEP} (\chi \rightarrow \tilde{\chi}) = \mathbb{E}_H \left\{ Q \left( \sqrt{\frac{E_S}{4N_0}} \gamma_{\chi \rightarrow \tilde{\chi}} (H) \right) \right\} \]

\[ = \left( \frac{1}{\pi} \int_{0}^{\pi/2} \mathbb{E}_H \left\{ \exp \left( - \frac{E_S}{8N_0 \sin^2 (\theta)} \gamma_{\chi \rightarrow \tilde{\chi}} (H) \right) \right\} d\theta \right) \]

\[ = \left( \frac{1}{\pi} \int_{0}^{\pi/2} \mathcal{L}_{\gamma_{\chi \rightarrow \tilde{\chi}}} \left( \frac{E_S}{8N_0 \sin^2 (\theta)} \right) d\theta \right) \]

\[ \text{APEP} (\chi \rightarrow \tilde{\chi}) \]

\[ = \frac{1}{\pi} \int_{0}^{\pi/2} \prod_{\tilde{\lambda}_q \in \mathcal{C}(R_n \Psi(\chi, \tilde{\chi}))} \left( 1 + \frac{E_S}{8N_0 \sin^2 (\theta)} \tilde{\lambda}_q \right)^{-1} d\theta \]

\[ \xrightarrow{E_S/N_0 \gg 1} \left( G_c (\chi, \tilde{\chi}) \frac{E_S}{4N_0} \right)^{-G_c (\chi, \tilde{\chi})} \]

Spatially-Modulated Space-Time-Coded MIMO (7/23)

Description of Transmission Modes

<table>
<thead>
<tr>
<th>Mode</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>GSSK (Generalized Space Shift Keying)</td>
<td>The spatial–constellation diagram is chosen such that its symbols (a^{(\alpha)} \in A) have (N_\alpha = N_{\alpha'}) for (\alpha = 1, 2, \ldots, A) and (A = 2 \left\lfloor \log_2 \left( \frac{N_t}{N_{\alpha'}} \right) \right\rfloor). This implies (A \subseteq \mathcal{U}).</td>
</tr>
<tr>
<td>MRSSK (Max–Rate Space Shift Keying)</td>
<td>The spatial–constellation diagram is chosen such that it contains the largest number of symbols (a^{(\alpha)} \in A) with (N_\alpha) assuming all possible values in ([0, N_t]) for (\alpha = 1, 2, \ldots, A) and (A = 2^{N_t-1}). This implies (A = \mathcal{U}).</td>
</tr>
<tr>
<td>MRSSK(_0) (Max–Rate Space Shift Keying “without the all–zero symbol”)</td>
<td>The spatial–constellation diagram is chosen such that it contains the largest number of symbols (a^{(\alpha)} \in A) with (N_\alpha) except (a^{(\alpha)} = 0_{N_t}) with (N_\alpha) assuming all possible values in ((0, N_t]) for (\alpha = 1, 2, \ldots, A) and (A = 2^{N_t-1}). This implies (A \subseteq \mathcal{U}(_0)).</td>
</tr>
<tr>
<td>SMSTT (Spatially–Modulated Space–Time–Transmission)</td>
<td>The spatial–constellation diagram is chosen such that its symbols (a^{(\alpha)} \in A) have (N_\alpha = N_{\alpha'}) for (\alpha = 1, 2, \ldots, A) and (A = 2 \left\lfloor \log_2 \left( \frac{N_t}{N_{\alpha'}} \right) \right\rfloor). This implies (A \subseteq \mathcal{U}). The mother space–time code (M(\cdot)) is transmitted from the active antennas in (a^{(\alpha)} \in A) over (N_\alpha) time–slots.</td>
</tr>
<tr>
<td>Rand (Random)</td>
<td>The symbols of the spatial–constellation diagram (a^{(\alpha)} \in A \subseteq \mathcal{U}) are randomly chosen but kept fixed for the whole communication.</td>
</tr>
<tr>
<td>SetPart (Set Partitioning) and NuSetPart (Non–Uniform Set Partitioning)</td>
<td>The symbols of the spatial–constellation diagram (a^{(\alpha)} \in A \subseteq \mathcal{U}) are chosen such that the active antennas in each symbol result in a partition of the antenna–array. Once chosen, the symbols are kept fixed for the whole communication. SetPart is obtained if (N_\alpha = N_{\alpha'}) for (\alpha = 1, 2, \ldots, A). NuSetPart is obtained if (N_\alpha) can be different for (\alpha = 1, 2, \ldots, A).</td>
</tr>
<tr>
<td>ISF (Identical Shaping Filters)</td>
<td>The shaping filters are the same for all the antenna–elements, i.e., (w_t(\xi) = w_0(\xi)) for (t = 1, 2, \ldots, N_t).</td>
</tr>
<tr>
<td>OSF (Orthogonal Shaping Filters)</td>
<td>The shaping filters are different and time–orthogonal for all the antenna–elements, i.e., (\int_0^{T_\alpha} w_{t_1}(\xi) w_{t_2}(\xi) d\xi = 0) for (t_1 \neq t_2 = 1, 2, \ldots, N_t).</td>
</tr>
<tr>
<td>SWOSF (Symbol–Wise Orthogonal Shaping Filters)</td>
<td>Let (a^{(\alpha_1)} \in A) and (a^{(\alpha_2)} \in A) be two generic symbols of the spatial–constellation diagram for (\alpha_1 \neq \alpha_2 = 1, 2, \ldots, A). Let (w_\alpha(\cdot)) be the common shaping filter used by all active antennas in (a^{(\alpha)}). Then the shaping filters are chosen such that (\int_0^{T_\alpha} w_{\alpha_1}(\xi) w_{\alpha_2}(\xi) d\xi = 0). Unlike OSF, the active antennas in (a^{(\alpha)}) have the same shaping filter.</td>
</tr>
</tbody>
</table>
### Acronym of Transmission Modes, Rate (R), Diversity (D), and Examples (X=\{ISF, OSF\}, Y=\{ISF, OSF, SWOSF\})

- **TM1–GSSK–Rand–X(N_t, N_\alpha):** \( R = \lceil \log_2 \left( \frac{N_t}{N_\alpha} \right) \rceil \) and \( D = N_r \) if \( X=\text{ISF} \), \( D = 2N_r \) if \( X=\text{OSF} \)
  
  **Example:** \( \mathcal{A} = \left\{ [0, 0, 1, 1]^T, [0, 1, 1, 0]^T, [1, 1, 0, 0]^T, [1, 0, 0, 1]^T \right\} \)

- **TM1–GSSK–SetPart–Y(N_t, N_\alpha):** \( R = \log_2 \left( \left\lfloor \frac{N_t}{N_\alpha} \right\rfloor \right) \) and \( D = N_r \) if \( Y=\text{ISF} \), \( D = 2N_\alpha N_r \) if \( Y=\text{OSF} \), \( D = 2N_r \) if \( Y=\text{SWOSF} \)
  
  **Example:** \( \mathcal{A} = \left\{ [0, 0, 1, 1]^T, [1, 1, 0, 0]^T \right\} \)

- **TM1–NuSetPart–Y(N_t, N_\alpha):** \( R = \log_2 (A) \) and \( D = N_r \) if \( Y=\text{ISF} \), \( D = \left[ \min \{N_\alpha\} + \min \{N_\alpha - \min \{N_\alpha\}\} \right] N_r \) if \( Y=\text{OSF} \), \( D = 2N_r \) if \( Y=\text{SWOSF} \)
  
  **Example:** \( \mathcal{A} = \left\{ [1, 1, 1, 0]^T, [0, 0, 0, 1]^T \right\} \) for \( N_\alpha = \{3, 1\} \)

- **TM1–MRSSK–X(N_t):** \( R = N_t \) and \( D = N_r \)
  
  **Example:** \( \mathcal{A} = \mathcal{U} \)

- **TM1–MRSSK\_0–Rand–X(N_t):** \( R = N_t - 1 \) and \( D = N_r \) if \( X=\text{ISF} \), \( D = 2N_r \) if \( X=\text{OSF} \)
  
  **Example:** \( \mathcal{A} = \left\{ [0, 0, 0, 1]^T, [0, 0, 1, 1]^T, [0, 1, 1, 0]^T, [0, 1, 1, 1]^T, [1, 1, 0, 0]^T, [1, 1, 0, 1]^T, [1, 1, 1, 0]^T \right\} \)

- **TM2–SMSTT–Rand–X(N_t, N_\alpha):** \( R = \left\lfloor \frac{1}{N_s} \right\rfloor \lceil \log_2 \left( \frac{N_t}{N_\alpha} \right) \rceil \) \( + (N_M/N_s) \log_2 (M) \) and \( D = N_r \) if \( X=\text{ISF} \), \( D \geq N_d N_r \) if \( X=\text{OSF} \)
  
  **Example:** \( \mathcal{A} = \left\{ [0, 0, 1, 1]^T, [0, 1, 1, 0]^T, [1, 1, 0, 0]^T, [1, 0, 0, 1]^T \right\} \) and \( M(\cdot) = H2 \)

- **TM2–SMSTT–SetPart–Y(N_t, N_\alpha):** \( R = \left\lfloor \frac{1}{N_s} \right\rfloor \log_2 \left( \left\lfloor \frac{N_t}{N_\alpha} \right\rfloor \right) \) \( + (N_M/N_s) \log_2 (M) \) and \( D = N_\alpha N_r \)
  
  **Example:** \( \mathcal{A} = \left\{ [0, 0, 1, 1]^T, [1, 1, 0, 0]^T \right\} \) and \( M(\cdot) = H2 \)
Example 1

$N_t=4$, $N_a=2$, $R=1$, $\text{Div}=4$ ($2*N_a*N_r$)

Example 2

Same Shaping Filters at Tx

- This scheme achieves transmit-diversity equal to 2 ($N_a * Nr$) but multi-stream decoding is needed.

Example 3

Time-Orthogonal Shaping Filters at Tx

- This scheme achieves transmit-diversity equal to 2 \((N_a*N_r)\) with single-stream decoding


\[
\Lambda (\bar{\alpha}, \bar{\mu}) \propto \sum_{r=1}^{N_r} \left\{ \left( \frac{E_S}{2} \right) |H_{r,\bar{i}_1}|^2 |\bar{\mu}_1|^2 + \left( \frac{E_S}{2} \right) |H_{r,\bar{i}_2}|^2 |\bar{\mu}_1|^2 - 2\text{Re} \left\{ \sqrt{\frac{E_S}{2}} H_{r,\bar{i}_2}^* \bar{\mu}_1 \bar{z}_{1,r}^{(\bar{i}_1)} \right\} - 2\text{Re} \left\{ \sqrt{\frac{E_S}{2}} H_{r,\bar{i}_2}^* \bar{\mu}_1 \bar{z}_{2,r}^{(\bar{i}_2)} \right\} \right\}
\]

\[
\Lambda_1 (\bar{\alpha}, \bar{\mu}_1)
\]

\[
\Lambda_2 (\bar{\alpha}, \bar{\mu}_2)
\]

\[
\left\{ \Lambda_1 (\bar{\alpha}, \bar{\mu}_1) = \sum_{r=1}^{N_r} \left[ \left( \frac{E_S}{2} \right) |H_{r,\bar{i}_1}|^2 |\bar{\mu}_1|^2 + \left( \frac{E_S}{2} \right) |H_{r,\bar{i}_2}|^2 |\bar{\mu}_1|^2 - 2\sqrt{E_S/2} \text{Re} \left\{ H_{r,\bar{i}_1}^* H_{r,\bar{i}_2} \bar{\mu}_1 \bar{\mu}_1 \rho((t_1, \bar{i}_1)) + H_{r,\bar{i}_2}^* H_{r,\bar{i}_1} \bar{\mu}_1 \bar{\mu}_1 \rho((t_2, \bar{i}_2)) \right\} \right\} = \mathcal{I}_1 (\bar{\alpha}, \bar{\mu}_1)
\]

\[
\left\{ \Lambda_2 (\bar{\alpha}, \bar{\mu}_2) = \sum_{r=1}^{N_r} \left[ \left( \frac{E_S}{2} \right) |H_{r,\bar{i}_1}|^2 |\bar{\mu}_2|^2 + \left( \frac{E_S}{2} \right) |H_{r,\bar{i}_2}|^2 |\bar{\mu}_2|^2 - 2\sqrt{E_S/2} \text{Re} \left\{ H_{r,\bar{i}_1}^* H_{r,\bar{i}_2} \bar{\mu}_2 \bar{\mu}_2 \rho((t_1, \bar{i}_1)) + H_{r,\bar{i}_2}^* H_{r,\bar{i}_1} \bar{\mu}_2 \bar{\mu}_2 \rho((t_2, \bar{i}_2)) \right\} \right\} = \mathcal{I}_2 (\bar{\alpha}, \bar{\mu}_2)
\]

\[
\left\{ \mathcal{I}_1 (\bar{\alpha}, \bar{\mu}_1) = -2\sqrt{E_S/2} \text{Re} \left\{ H_{r,\bar{i}_2}^* H_{r,\bar{i}_1}^* \bar{\mu}_1 \bar{\mu}_2 \rho((t_1, \bar{i}_2)) \right\}
\right. 
\]

\[
+ 2\sqrt{E_S/2} \text{Re} \left\{ H_{r,\bar{i}_1}^* H_{r,\bar{i}_2} \bar{\mu}_1 \bar{\mu}_2 \rho((t_2, \bar{i}_1)) \right\}
\]

\[
\right. 
\left. \mathcal{I}_2 (\bar{\alpha}, \bar{\mu}_2) = 2\sqrt{E_S/2} \text{Re} \left\{ H_{r,\bar{i}_1}^* H_{r,\bar{i}_2} \bar{\mu}_1 \bar{\mu}_2 \rho((t_1, \bar{i}_2)) \right\}
\right.
\]

\[
- 2\sqrt{E_S/2} \text{Re} \left\{ H_{r,\bar{i}_1}^* H_{r,\bar{i}_2} \bar{\mu}_1 \bar{\mu}_1 \rho((t_2, \bar{i}_1)) \right\}
\]

\[
\mathcal{I}_2 (\bar{\alpha}, \bar{\mu}_2) = 0
\]

\[
\mathcal{I}_1 (\bar{\alpha}, \bar{\mu}_1) = 0
\]

\[ (\hat{\alpha}, \hat{\mu}_1, \hat{\mu}_2) = \arg \min_{\hat{\alpha} \in \mathcal{A}, \mu_1 \in \mathcal{M}, \mu_2 \in \mathcal{M}} \{ \Lambda_1 (\hat{\alpha}, \hat{\mu}_1) + \Lambda_2 (\hat{\alpha}, \hat{\mu}_2) \} \]

\[ \overset{(1)}{=} \arg \min_{\hat{\alpha} \in \mathcal{A}} \left\{ \min_{\mu_1 \in \mathcal{M}, \mu_2 \in \mathcal{M}} \{ \Lambda_1 (\hat{\alpha}, \hat{\mu}_1) + \Lambda_2 (\hat{\alpha}, \hat{\mu}_2) \} \right\} \]

\[ \overset{(2)}{=} \arg \min_{\hat{\alpha} \in \mathcal{A}} \left\{ \min_{\mu_1 \in \mathcal{M}} \{ \Lambda_1 (\hat{\alpha}, \hat{\mu}_1) \} + \min_{\mu_2 \in \mathcal{M}} \{ \Lambda_2 (\hat{\alpha}, \hat{\mu}_2) \} \right\} \]

Step 1: For every hypothesis $a^{(\tilde{\alpha})} \in A$ of the spatial-constellation diagram, compute:

$$\hat{\mu}_1 (\tilde{\alpha}) = \arg\min_{\tilde{\mu}_1 \in \mathcal{M}} \{ \Lambda_1 (\tilde{\alpha}, \tilde{\mu}_1) \}$$

$$\hat{\mu}_2 (\tilde{\alpha}) = \arg\min_{\tilde{\mu}_2 \in \mathcal{M}} \{ \Lambda_2 (\tilde{\alpha}, \tilde{\mu}_2) \}$$

Step 2: Compute the estimate of the symbol belonging to the spatial-constellation diagram as follows:

$$\hat{\alpha} = \arg\min_{a^{(\tilde{\alpha})} \in A} \{ \Lambda_1 (\tilde{\alpha}, \hat{\mu}_1 (\tilde{\alpha})) + \Lambda_2 (\tilde{\alpha}, \hat{\mu}_2 (\tilde{\alpha})) \}$$

Step 3: Compute the estimates of the two symbols belonging to the signal-constellation diagram as follows:

$$\hat{\mu}_1 = \hat{\mu}_1 (\hat{\alpha}) \quad \text{and} \quad \hat{\mu}_2 = \hat{\mu}_2 (\hat{\alpha})$$


\[
\hat{\mu}_m |_{m=1,2,\ldots,N_M} = \arg\min_{\hat{\mu}_m \in \mathcal{M}} \left\{ \Lambda_m (\hat{\alpha}, \hat{\mu}_m) = \sum_{r=1}^{N_r} \left[ \frac{E_S}{2} \left( \sum_{\tau=1}^{N_t} |H_r,_{\bar{\tau}}|^2 \right) |\hat{\mu}_m|^2 - 2\sqrt{E_S/2} \Re \left\{ T_m^{(\text{HX})} (t, \bar{\tau}; r) \hat{\mu}_m^* \right\} \right] \right\}
\]

\[
\hat{\alpha} = \arg\min_{\alpha(\bar{\alpha}) \in \mathcal{A}} \left\{ \sum_{m=1}^{N_M} \Lambda_m (\hat{\alpha}, \hat{\mu}_m (\hat{\alpha})) \right\}
\]

\[
\hat{\mu}_m |_{m=1,2,\ldots,N_M} = \hat{\mu}_m (\hat{\alpha} = \hat{\alpha}) |_{m=1,2,\ldots,N_M}
\]

Example: OSTBC Tarokh-H3

\[
\begin{align*}
\gamma_1^{(\text{H3})} (t, \bar{\tau}; r) &= z_{1,r}^{(i_1)} H_{r,_{\bar{\tau}}}^* + \left( z_{2,r}^{(i_2)} \right)^* H_{r,_{\bar{\tau}}} + (1/2) \left( z_{4,r}^{(i_3)} - z_{3,r}^{(i_3)} \right) H_{r,_{\bar{\tau}}}^* - (1/2) \left( z_{3,r}^{(i3)} + z_{4,r}^{(i3)} \right)^* H_{r,_{\bar{\tau}}} \\
\gamma_2^{(\text{H3})} (t, \bar{\tau}; r) &= z_{1,r}^{(i_2)} H_{r,_{\bar{\tau}}}^* - \left( z_{2,r}^{(i_1)} \right)^* H_{r,_{\bar{\tau}}} + (1/2) \left( z_{3,r}^{(i_3)} + z_{4,r}^{(i_3)} \right) H_{r,_{\bar{\tau}}}^* - (1/2) \left( z_{4,r}^{(i3)} - z_{3,r}^{(i3)} \right)^* H_{r,_{\bar{\tau}}} \\
\gamma_3^{(\text{H3})} (t, \bar{\tau}; r) &= \left( 1/\sqrt{2} \right) \left( z_{1,r}^{(i_1)} + z_{2,r}^{(i_3)} \right) H_{r,_{\bar{\tau}}}^* + \left( 1/\sqrt{2} \right) \left( z_{3,r}^{(i_1)} \right)^* H_{r,_{\bar{\tau}}} + \left( 1/\sqrt{2} \right) \left( z_{3,r}^{(i_2)} \right)^* H_{r,_{\bar{\tau}}} \\
&\quad + \left( 1/\sqrt{2} \right) \left( z_{4,r}^{(i_2)} \right)^* H_{r,_{\bar{\tau}}} - \left( 1/\sqrt{2} \right) \left( z_{4,r}^{(i_3)} \right)^* H_{r,_{\bar{\tau}}}
\end{align*}
\]

Spatially-Modulated Space-Time-Coded MIMO (13/23)

Diversity Analysis ($N_r = 1 - R = 4$ bpcu)

![Graph showing ABEP vs. $E_s/N_0$ for different modulation schemes.](image)
Diversity Analysis ($N_r = 2 - R = 4$ bpcu)
Multi vs. Single-Stream Decoding ($R = 4$ bpcu)
\( N_r = 1 \)
\( R = 4 \) bpcu
\( N_r = 1 \)
\( R = 6 \text{ bpcu} \)
$N_r = 2$

$R = 6$ bpcu
$N_r = 4$

$R = 6$ bpcu
$N_r = 1$

$R = 8 \text{ bpcu}$
Spatially-Modulated Space-Time-Coded MIMO (21/23)

\[ N_r = 2 \]
\[ R = 8 \text{ bpcu} \]
$N_r = 4$

$R = 8 \text{ bpcu}$
OSF-MIMO
$N_r = 2$
$R = 8 \text{ bpcu}$
Spatially Modulated Orthogonal Space-Time Block Codes with Non-Vanishing Determinants

Minh-Tuan Le, Vu-Duc Ngo, Hong-Anh Mai, Xuan Nam Tran, and Marco Di Renzo, Member, IEEE

Abstract—This paper proposes a multiple-input multiple-output (MIMO) transmission scheme for \( M \)-ary modulations, called Spatially Modulated Orthogonal Space-Time Block Coding (SM-OSTBC), based on the concept of Spatial Constellation (SC) high multiplexing gain by transmitting data streams in parallel from different transmit antennas, yet at the cost of a significant decoding complexity that is required for reducing the impact of Inter-Channel Interference (ICI). Orthogonal Space-Time

\[
\text{Rate} = N_t + \log_2(M) - 2
\]

Fig. 1. Block diagram of the SM-OSTBC transmitter.
Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
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13. Relay-Aided SM
14. SM in Heterogeneous Cellular Networks
15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
Multi-Hop Networks:

- **Advantages**: better performance, extended coverage...
- **Disadvantages**: additional resources (relays, time-slots, frequencies), capacity reduction, half-duplex constraint...
Cooperative Networks:

- **Advantages**: better performance, (macro) diversity…
- **Disadvantages**: additional resources (relays, time-slots, frequencies), capacity reduction, half-duplex constraint…
Dual-Hop Spatial Modulation

\[ P_b(\tilde{\gamma}_{sd}) = P_b(\tilde{\gamma}_{sr}) + P_b(\tilde{\gamma}_{rd}) - 2P_b(\tilde{\gamma}_{sr})P_b(\tilde{\gamma}_{rd}) \]

Relay-Aided SM (4/24)

$N_r^F = 2$, $N_t^F = 2$, $N_r^d = 2$ and 7 bits/s/Hz per link spectral efficiency

- Bit-Error-Ratio vs SNR / dB
- DF -- An(s, d)
- DF -- Sim(s, d)
- Dh–SM -- An(s, d) $N_t^S = 32$
- Dh–SM -- Sim(s, d) $N_t^S = 2$
- Dh–SM -- Sim(s, d) $N_t^S = 4$
- Dh–SM -- Sim(s, d) $N_t^S = 32$
Relay-Aided SM (5/24)

$N_t^s = 32$, $N_r^f = 2$, $N_{r}^{d} = 2$ and 7 bits/s/Hz per link spectral efficiency

[Graph showing bit-error ratio vs. SNR for different scenarios with varying $N_t^s$]
Relay-Aided SM (6/24)

\[ N_s^t = 2, \quad N_r^f = 2 \] and 7 bits/s/Hz per link spectral efficiency

![Graph showing bit-error ratio vs. SNR in dB for different conditions.](image)
Relay-Aided SM (7/24)

![Graph showing Bit Error Ratio vs. SNR for different configurations of relay-aided SM systems. The graph illustrates the performance of different modulation techniques and shows the impact of varying parameters on spectral efficiency.](image-url)
Relay-Aided SM (8/24)
Virtual SM-MIMO for the Uplink

- In TS-1, MS broadcasts its own info symbol to a group of $N_R$ relays. Each symbol has $\log_2(N_R)$ bits (QAM or PSK).
- The relays decode the received symbol without any coordination among them.
- Each relay is assigned an individual ID. If the symbol received from MS coincides with the ID, then the relay is activated for transmission.
- Thus, the relays play the role of a distributed spatial constellation diagram.
- The relay-activation process conveys information.
- Errors may occur, and so multiple or no relays may wake up.
Virtual SM-MIMO for the Uplink

\[
\hat{b}^{(D)}_S = \arg\min_{(\hat{b}^{(D)}_S, a^{(R_1)}_{tx}, a^{(R_2)}_{tx})} \left\{ \sum_{d=1}^{2} |y_{RD \bar{d}} - \mathcal{F}\left(\hat{b}^{(D)}_S, a^{(R_1)}_{tx}, a^{(R_2)}_{tx}\right)|^2 \right\}
\]

\[
\mathcal{F}\left(\hat{b}^{(D)}_S = 0, a^{(R_1)}_{tx} = 1, a^{(R_2)}_{tx} = 0\right) = \sqrt{E_m h_{R_1 D \bar{d}}}
\]

\[
\mathcal{F}\left(\hat{b}^{(D)}_S = 1, a^{(R_1)}_{tx} = 0, a^{(R_2)}_{tx} = 1\right) = \sqrt{E_m h_{R_2 D \bar{d}}}
\]

Relay-Aided SM (11/24)

Optimal (Error-Aware) Demodulator

\[
\begin{align*}
    y_{SR_1} &= h_{SR_1}x_S + n_{SR_1} \\
    y_{SR_2} &= h_{SR_2}x_S + n_{SR_2}
\end{align*}
\]

\[
\begin{align*}
    \hat{b}_S^{(R_1)} &= \underset{\hat{b}_S \in \{0,1\}}{\text{arg min}} \left\{ \left| y_{SR_1} - \sqrt{E_m}h_{SR_1} \left( 1 - 2\hat{b}_S \right) \right|^2 \right\} \\
    \hat{b}_S^{(R_2)} &= \underset{\hat{b}_S \in \{0,1\}}{\text{arg min}} \left\{ \left| y_{SR_2} - \sqrt{E_m}h_{SR_2} \left( 1 - 2\hat{b}_S \right) \right|^2 \right\}
\end{align*}
\]

\[
y_{RD_\bar{d}} = \sqrt{E_m}h_{R_1D_\bar{d}} \Delta \left( \text{ID}_{R_1}, \hat{b}_S^{(R_1)} \right) + \sqrt{E_m}h_{R_2D_\bar{d}} \Delta \left( \text{ID}_{R_2}, \hat{b}_S^{(R_2)} \right) + n_{RD_\bar{d}}
\]

\[
\Delta \left( \text{ID}_{R_\bar{r}}, \hat{b}_S^{(R_\bar{r})} \right) = \begin{cases} 
1 & \text{if } \text{ID}_{R_\bar{r}} = \hat{b}_S^{(R_\bar{r})} \\
0 & \text{if } \text{ID}_{R_\bar{r}} \neq \hat{b}_S^{(R_\bar{r})}
\end{cases}
\]

Relay-Aided SM (12/24)

Optimal (Error-Aware) Demodulator

\[
\hat{b}_S^{(D)} = \arg \max_{\hat{b}_S^{(D)} \in \{0,1\}} \left\{ \sum_{a_{t_x}^{(R_1)} = 0}^{1} \sum_{a_{t_x}^{(R_2)} = 0}^{1} \left\{ \prod_{d=1}^{2} \mathcal{P} \{ y_{RD_d} \mid \mathcal{H} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_1)}, a_{t_x}^{(R_2)} \right) \} \right\} \mathcal{P} \{ \mathcal{H} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_1)}, a_{t_x}^{(R_2)} \right) \} \right\}
\]

\[
\mathcal{P} \{ y_{RD_d} \mid \mathcal{H} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_1)}, a_{t_x}^{(R_2)} \right) \} = \exp \left( -\frac{\left| y_{RD_d} - \left( \sqrt{E_m} h_{R_1 D_d} a_{t_x}^{(R_1)} + \sqrt{E_m} h_{R_1 D_d} a_{t_x}^{(R_2)} \right) \right|^2}{N_0} \right)
\]

\[
\mathcal{P} \{ \mathcal{H} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_1)}, a_{t_x}^{(R_2)} \right) \} = w_{SR_1} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_1)} \right) w_{SR_2} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_2)} \right)
\]

\[
w_{SR_r} \left( \hat{b}_S^{(D)}, a_{t_x}^{(R_r)} \right) = \begin{cases} 
1 - Q \left( \sqrt{2|h_{SR_r}|^2 (E_m / N_0)} \right) & \text{if } \left( \hat{b}_S^{(D)} = \bar{r} - 1 \text{ and } a_{t_x}^{(R_r)} \neq 0 \right) \text{ or } \left( \hat{b}_S^{(D)} \neq \bar{r} - 1 \text{ and } a_{t_x}^{(R_r)} = 0 \right) \\
Q \left( \sqrt{2|h_{SR_r}|^2 (E_m / N_0)} \right) & \text{otherwise}
\end{cases}
\]

Relay-Aided SM (13/24)
Relay-Aided SM (14/24)

Spectral-Efficient Relaying

**Repetition Relaying**

| MS(MS) → Rx | R1(MS) → BS | R2(MS) → BS | R1(R1) → BS | R2(R2) → BS |

**Selective Relaying**

| MS(MS) → Rx | Rbest(MS) → BS | R1(R1) → BS | R2(R2) → BS |

**Network Coding (NC) Based - Phoenix**

| MS(MS) → Rx | R1(MS,R1) → BS | R2(MS,R2) → BS |

**DSTBC Relaying – Alamouti Based**

| MS(MS1) → Rx | R1(MS1) → BS | R1(-MS2*) → BS | R2(MS2) → BS |
| MS(MS2) → Rx | R2(MS2) → BS | R2(MS1*) → BS |

**Spatial Modulation Based**

| MS(MSi) → Rx | id=MS1 Rnid is silent id=MS2 Rnid is silent |

A new relaying protocol based on Spatial Modulation (the Relays have data in their buffers)
Distributed SM

\[
\begin{aligned}
    y_{SR_r} &= \sqrt{E_S} h_{SR_r} s_{tx} + n_{SR_r} \\
    y_{SD} &= \sqrt{E_S} h_{SD} s_{tx} + n_{SD}
\end{aligned}
\]

\[
\hat{s}_{tx}^{(R_r)} = \arg \min_{p_m \in \{p_1, p_2, \ldots, p_M\}} \left\{ \left| y_{SR_r} - \sqrt{E_S} h_{SR_r} p_m \right|^2 \right\}
\]

\[
\hat{x}_S^{(R_r)} = M_S^{-1} \left( \hat{s}_{tx}^{(R_r)} \right)
\]

\[
r_{tx}^{(R_r)} = \begin{cases} 
    M_R (x_{R_r}) & \text{if } \textbf{ID}_{R_r} = \hat{x}_S^{(R_r)} \\
    0 & \text{if } \textbf{ID}_{R_r} \neq \hat{x}_S^{(R_r)}
\end{cases}
\]

\[
y_{RD} = \sum_{r=1}^{M} \left( \sqrt{E_{R_r}} h_{R_r} D r_{tx}^{(R_r)} \right) + n_{RD}
\]

Relay-Aided SM (16/24)

Optimal (Error-Aware) Demodulator

\[
\hat{x}_S^{(D)}, \hat{x}_{R_1}^{(D)}, \hat{x}_{R_2}^{(D)} = \arg \max_{x_S^{(D)} \in \{0,1\}, x_{R_1}^{(D)} \in \{0,1,\mathcal{N}\}, x_{R_2}^{(D)} \in \{0,1,\mathcal{N}\}} \{ \mathcal{P} \{ [y_{SD}, y_{RD}] | \mathcal{H}(x_S^{(D)}, x_{R_1}^{(D)}, x_{R_2}^{(D)}) \} \} \}
\]

\[
\mathcal{P} \{ [y_{SD}, y_{RD}] | \mathcal{H}(x_S^{(D)}, x_{R_1}^{(D)}, x_{R_2}^{(D)}) \} = \exp \left( -\frac{|y_{SD} - \sqrt{E_s} h_{SD} m_S (x_S^{(D)})|^2}{N_0} \right) \times \exp \left( -\frac{|y_{RD} - \sqrt{E_{R_1} h_{R_1 D} m_{R_1} (x_{R_1}^{(D)})} + \sqrt{E_{R_2} h_{R_2 D} m_{R_2} (x_{R_2}^{(D)})}|^2}{N_0} \right)
\]

\[
\omega_{SR_r}(x_S^{(D)}, x_{R_r}^{(D)}) = \begin{cases} 
1 - Q \left( \sqrt{2|h_{SR_r}|^2 (E_s/N_0)} \right) & \text{if } (x_S^{(D)} = r - 1 \text{ and } x_{R_r}^{(D)} \neq \mathcal{N}) \text{ or } (x_S^{(D)} \neq r - 1 \text{ and } x_{R_r}^{(D)} = \mathcal{N}) \\
Q \left( \sqrt{2|h_{SR_r}|^2 (E_s/N_0)} \right) & \text{otherwise}
\end{cases}
\]

Relay-Aided SM (17/24)

Diversity order of the source is 2 (analytically proved)
Relay-Aided SM (18/24)

$$y_{RD}^{\text{SPM}} = \sqrt{E_{R_1}} h_{R_1D} \left( \sqrt{1 - \gamma^2 M_R (x_{R_1})} + \gamma M_S \left( \hat{x}_{S}^{(R_1)} \right) \right) + n_{RD}$$
Relay-Aided SM (19/24)
Decoding-and-Forward (DF) Non-Orthogonal Relaying

Relaying Phase
- \( x = [x_d, x_c] \): received from the source
- \( x_d \): spatial-constellation diagram
- \( x_c \): signal-constellation diagram

Capacity complementary cumulative distribution function (CCDF) comparison among:
- The general IGT scheme (general IGT)
- The specific IGT case with single-relay selection (SR-IGT)
- The benchmark in [*]
  (a) $M = 2$ relay nodes
  (b) $M = 4$ relay nodes

Fig. 6. Ergodic capacity vs. SNR comparisons among the general IGT scheme (general IGT), the specific IGT case with single-relay selection (SR-IGT), and the benchmark.
Relay-Aided SM (24/24)
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SM in Heterogeneous Cellular Networks (1/22)
Heterogeneous cellular systems are networks with different types of cells providing different QoS requirements to the users, which coexist and contend the wireless medium (macro, pico, femto, relays, DAEs, cognitive radios, etc.)

Thus, interference should be properly managed and/or exploited for reliable communications and energy efficiency.

Overlaid multi-tier heterogeneous scenario
SM in Heterogeneous Cellular Networks (3/22)

...what cellular will migrate to (Prof. Jeff Andrews, UT Austin)...

Traditional grid model
Actual 4G network today
Completely random BSs

Zoom w/ femtocells
Zoom w/ picocells too
Conventional approaches for the analysis and design of (heterogeneous) cellular networks (abstraction models) are:

- The Wyner model
- The single-cell interfering model
- The regular hexagonal or square grid model

However, these abstraction models:

- Are over-simplistic and/or inaccurate
- Require intensive numerical simulations and/or integrations
- Provide information only for specific BSs deployments
- No closed-form solutions and/or insights


SM in Heterogeneous Cellular Networks (5/22)

An Emerging (Tractable) Approach

- **RANDOM SPATIAL MODEL** for Heterogeneous Cellular Networks (HCNs):
  - K-tier network with BS locations modeled as independent marked Poisson Point Processes (PPPs)
  - PPP model is surprisingly good for 1-tier as well (macro BSs): lower bound to reality and trends still hold
  - PPP makes even more sense for HCNs due to less regular BSs placements for lower tiers (femto, etc.)

**Stochastic Geometry**
emerges as an effective tool for analysis, design, and optimization of HCNs
How It Works (Downlink – 1-tier)

- **Probe mobile terminal**
- **PPP-distributed macro base station**
How It Works (Downlink – 1-tier)

- Probe mobile terminal
- PPP-distributed macro base station
SM in Heterogeneous Cellular Networks (8/22)

How It Works (Downlink – 1-tier)

- Probe mobile terminal
- PPP-distributed macro base station

Useful link
How It Works (Downlink – 1-tier)

- Probe mobile terminal
- PPP-distributed macro base station

Useful link
How It Works (Downlink – 2-tier)


Base station distribution in Taipei City, Taiwan, shown on Google Map. Blue Δ’s are the locations of base stations.


Open source project OpenCellID: http://www.opencellid.org/
PPP better than (or same accuracy as) Hexagonal


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PPP better than (or same accuracy as) Hexagonal


Open source project OpenCellID: http://www.opencellid.org/
Preliminary Reference Scenario

- Probe mobile terminal
- PPP-distributed interfering lower-tier (e.g., femto) base stations
- Tagged macro base station at a fixed distance → cell association is neglected
SM in Heterogeneous Cellular Networks (17/22)

A Key Result from Stochastic Geometry and PPP Theory

\[
\Lambda \propto |\Delta_0|^2 U + 2 \text{Re}\{\Delta_0^*N}\} + 2 \text{Re}\{\Delta_0^*I_{\text{AGG}}\}
\]

Decision Metric
Useful Signal
AWGN
Aggregate Interference

\[
\bar{I}_{\text{AGG}} = \sum_{i \in \Phi_{\text{PPP}}} \left( \frac{\bar{Z}_i}{d_i^{b_I}} \right) \Rightarrow \bar{I}_{\text{AGG}} = B_I^{1/2}\bar{G}_I \sim S\alpha S(\alpha_I = 2/b_I, \gamma_I)
\]

\[
\begin{cases}
B_I \sim S\left(1/b_I, 1, \cos^{b_I}(\pi/2b_I)\right)
\end{cases}
\]

\[
M_{B_I}(s) = E_{B_I}\{\exp(-sB_I)\} = \exp\left(-s^{1/b_I}\right)
\]

\[
\bar{G}_I \sim CN\left(0, 4\gamma_I^{b_I}\right)
\]
**SM in Heterogeneous Cellular Networks (18/22)**

**Equivalent AWGN Channel**

\[
\Delta \propto \left| \Delta_0 \right|^2 U + 2 \text{Re} \left\{ \Delta_0^* \bar{N} \right\} + 2 \text{Re} \left\{ \Delta_0^* \left( B_I^{1/2} \bar{G}_I \right) \right\}
\]

- **Decision Metric**
- **Useful Signal**
- **AWGN**
- **Aggregate Interference**
- **Equivalent AWGN conditioning upon** \( B_I \)

**STEP 1:** The frameworks developed without interference can be applied by conditioning upon \( B_I \)

**STEP 2:** The conditioning can be removed either numerically or analytically (preferred)
The Bottom Line

- Closed-form results in **STEP 1** can be obtained from:
  
  \[ \text{ABEP}(B_I) \leq \text{ABEP}_{\text{signal}}(B_I) + \text{ABEP}_{\text{spatial}}(B_I) + \text{ABEP}_{\text{joint}}(B_I) \]

- The average over \( B_I \) in **STEP 2** can be computed using (e.g., for Nakagami-m fading):
  
  \[ J(\eta_p^2; \psi_p; \theta_p) = J(v_p) = \int_0^{\mu_p} E_{B_I} \left\{ \left( 1 + \frac{\chi_{B_I}^{(p)}}{\sin^2(\omega)} \right)^{-m_0N_r} \right\} d\omega \]
SM in Heterogeneous Cellular Networks (20/22)

Fig. 1. ASEP as a function of $E_P/N_0$, $N_t$ and $\lambda$. Setup: $\Omega_P = \Omega_I = 1$; $b = 4$; $X_0^2 = 1$; $N_r = 2$; $E_P = E_I$; PSK modulation with $M = 8$. The solid lines show the framework in (21), (23) and (26), the dashed lines the asymptotic framework in (29), and the markers the Monte Carlo simulations.
Fig. 2. ASEP as a function of $E_P/N_0$, $N_t$ and $\lambda$. Setup: $\Omega_P = \Omega_I = 1$; $b = 4$; $X_0^2 = 1$; $N_r = 2$; $E_I/N_0 = 40$dB; PSK modulation with $M = 16$. The solid lines show the framework in (21), (23) and (26), the dashed lines the asymptotic framework in (23), and the markers the Monte Carlo simulations.
Fig. 3. ASEP as a function of $E_P/N_0$, $N_r$ and $\lambda$. Setup: $\Omega_P = \Omega_I = 1$; $b = 4$; $X_0^2 = 1$; $N_t = 8$; $E_I/N_0 = 40$dB; PSK modulation with $M = 32$. The solid lines show the framework in (21), (23) and (26), the dashed lines the asymptotic framework in (30), and the markers the Monte Carlo simulations.
Outline

1. Introduction and Motivation behind SM-MIMO
2. History of SM Research and Research Groups Working on SM
3. Transmitter Design – Encoding
4. Receiver Design – Demodulation
5. Error Performance (Numerical Results and Main Trends)
6. Achievable Capacity
7. Channel State Information at the Transmitter
8. Imperfect Channel State Information at the Receiver
9. Multiple Access Interference
10. Energy Efficiency
11. Transmit-Diversity for SM
12. Spatially-Modulated Space-Time-Coded MIMO
13. Relay-Aided SM
14. SM in Heterogeneous Cellular Networks
15. SM for Visible Light Communications
16. Experimental Evaluation of SM
17. The Road Ahead – Open Research Challenges/Opportunities
18. Implementation Challenges of SM-MIMO
**HOW IT WORKS**
An overhead lamp fitted with an LED with signal-processing technology (below) streams data embedded in its beam at ultra-high speeds to the photo-detector. A receiver dongle then converts the tiny changes in amplitude into an electrical signal, which is then converted back into a data stream and transmitted to a computer or mobile device.
**SM for Visible Light Communications (4/13)**

<table>
<thead>
<tr>
<th>Attribute</th>
<th>RF wireless</th>
<th>Optical Wireless</th>
</tr>
</thead>
<tbody>
<tr>
<td>Link margin</td>
<td>Excellent, due to coherent detection and noise processes</td>
<td>Poor due to incoherent detection at receiver and noise mechanisms</td>
</tr>
<tr>
<td>Signal confinement</td>
<td>Typical propagation loss through walls: 1dB/cm</td>
<td>Confined by internal spaces</td>
</tr>
<tr>
<td>Regulation</td>
<td>Subject to licensing and regulation</td>
<td>Subject to eye safety regulation, currently no licensing</td>
</tr>
<tr>
<td>Availability of spectrum</td>
<td>Limited and expensive</td>
<td>1000s of THz of spectrum available</td>
</tr>
<tr>
<td>Ability to control radiation</td>
<td>Challenging owing to constraints on antenna size in most appliances</td>
<td>High degree of control using lenses and diffractive elements</td>
</tr>
<tr>
<td>Propagation</td>
<td>Scattering/Reflection and diffraction create broad area coverage</td>
<td>Ray propagation allows very tight confinement of radiation</td>
</tr>
<tr>
<td>Interface with fixed network</td>
<td>Need optoelectronic interface of directed RF to connect with fibre network</td>
<td>Possibility of using light from fiber network to create transparent interface</td>
</tr>
</tbody>
</table>

![Graph showing frequency bands](image)

Optical Wireless Setup and Channel

\[
h = \begin{cases} 
\frac{(k+1)A}{2\pi d^2} \cos^k \phi \cos \psi & 0 \leq \psi \leq \Psi_{1/2} \\
0 & \psi > \Psi_{1/2}
\end{cases}
\]

\[
k = \frac{-\ln(2)}{\ln(\cos \Phi_{1/2})}
\]

\(\Phi_{1/2} = 15^\circ\): Tx semi-angle

\(\Psi_{1/2} = 15^\circ\): Rx semi-angle

\(A = 1\text{cm}^2\): receiver detector area

$N_t = 8$, Rate = 5 bpcu
Fig. 4. Comparison of SM and RC for spectral efficiency of 2, 4 and 6 bit/s/Hz in 4 × 4 setup scenario (lines show simulation results and markers numerical ABEP results).
Fig. 6. Comparison of SM and RC for spectral efficiency of 4, 5 and 6 bit/s/Hz in 16×16 setup scenario (lines show simulation results and markers numerical ABEP results).
$N_t = N_r = 4$, Rate = 4 bpcu

\( N_t = N_r = 4, \text{ Rate } = 8 \text{ bpcu} \)

$N_t = N_r = 4$, Rate = 4, 8 bpcu, $d_{TX} = 0.7$

GSSK –VLC transmitter developed by the startup PureVLC
http://purevlc.com/
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Experimental Evaluation of SM (1/31)

- Performance assessment via channel measurements
  - Urban scenario (Bristol/UK) @ 2GHz carrier frequency

MIMO channel sounder

Post-processing

- Testbed implementation (NI-PXIe-1075 @ Heriot-Watt Univ. / UK)
  - Laboratory environment: 2x2 MIMO @ 2.3GHz carrier frequency
MIMO channel measurements are taken around the center of Bristol city (UK), using a MEDAV RUSK channel sounder.

The setup consists of a 4×4 MIMO, with 20 MHz bandwidth centered at 2 GHz.

The transmitter consists of a pair of dual polarized (±45°) Racal Xp651772 antennas separated by 2m, positioned atop a building, providing elevated coverage of central business and commercial districts of Bristol city.

A. Younis, W. Thompson, M. Di Renzo, C.-X. Wang, M. A. Beach, H. Haas, and P. M. Grant, "Performance of Spatial Modulation over Correlated and Uncorrelated Urban Channel Measurements", IEEE VTC-Fall, 2013. BEST PAPER AWARD
At the receiver, two different receiver devices are used, both equipped with four antennas:

- **A reference headset**, which is based on 4-dipoles mounted on a cycle helmet, thus avoiding any shadowing by the user.

- **A laptop**, which is equipped with 4 Printed Inverted F Antennas (PIFA) fitted inside the back of the display panel.

---

Channel Measurements

- 58 measurement locations are chosen around the city.
- At each location the user walked, holding the laptop in front of him and the reference device on his head, in a straight line roughly 6 m long, until 4096 channel snapshots were recorded.
- A second measurement is then taken with the user walking a second path perpendicular to the first.
- As the measurement speed is significantly faster than the coherence time of the channel, the measurements are averaged in groups of four to reduce measurement noise.
- One set of measurement results with the laptop and reference device, and a second set of only the reference device measurements taken at the same locations, but on different days.

A. Younis, W. Thompson, M. Di Renzo, C.-X. Wang, M. A. Beach, H. Haas, and P. M. Grant, "Performance of Spatial Modulation over Correlated and Uncorrelated Urban Channel Measurements", IEEE VTC-Fall, 2013. BEST PAPER AWARD
This provides a total of 348 different measurement sets, each containing 1024 snapshots of a 4×4 MIMO channel, with 128 frequency bins spanning the 20 MHz bandwidth.

As the simulations are carried out using flat fading channels, a single frequency bin centered around 2 GHz, is chosen from each measurement snapshot to create the narrowband channel.

Two MIMO test cases are investigated:

- “Small-scale” MIMO, which are the original 4x4 channel measurements
- “Large-scale” MIMO, where, by manipulating the original measurements, larger virtual MIMO systems are created

A. Younis, W. Thompson, M. Di Renzo, C.-X. Wang, M. A. Beach, H. Haas, and P. M. Grant, "Performance of Spatial Modulation over Correlated and Uncorrelated Urban Channel Measurements", IEEE VTC-Fall, 2013. BEST PAPER AWARD
Small-Scale MIMO

- For small-scale MIMO, locations whose channel taps experienced Rayleigh fading are used.
- The chi-squared goodness of fit test, with a significance level of 1%, is used to identify Rayleigh fading channels.
- 20 out of the 348 measurement sets (each containing 1024 snapshots), fulfilled this requirement and are kept for further processing.
- For each location the transmit and receive correlation matrices are estimated, then the decay of the correlation, based on the antenna indices, is fitted to an exponential decay model ($\gamma$ is the correlation decay coefficient):

$$
R = \begin{bmatrix}
1 & \exp(-\gamma) & \exp(-2\gamma) & \exp(-3\gamma) \\
\exp(-\gamma) & 1 & \exp(-\gamma) & \exp(-2\gamma) \\
\exp(-2\gamma) & \exp(-\gamma) & 1 & \exp(-\gamma) \\
\exp(-3\gamma) & \exp(-2\gamma) & \exp(-\gamma) & 1
\end{bmatrix}
$$

A. Younis, W. Thompson, M. Di Renzo, C.-X. Wang, M. A. Beach, H. Haas, and P. M. Grant, "Performance of Spatial Modulation over Correlated and Uncorrelated Urban Channel Measurements", IEEE VTC-Fall, 2013. BEST PAPER AWARD.
Correlated channels:

- Two measurement sets with the lowest mean square error between the model and the actual correlation matrices are retained. Both of them are from the laptop device.
- The measured decay coefficients for the transmitter and receiver are 0.41 and 0.99 for the first channel and 0.36 and 0.75 for the second channel, respectively.

Uncorrelated channels:

- The two measurement sets with the lowest average correlation coefficient are kept.
- One is from the laptop and the other from the reference device.

A. Younis, W. Thompson, M. Di Renzo, C.-X. Wang, M. A. Beach, H. Haas, and P. M. Grant, "Performance of Spatial Modulation over Correlated and Uncorrelated Urban Channel Measurements", IEEE VTC-Fall, 2013. BEST PAPER AWARD.
The following post-processing steps are used to create the large-scale channel measurements from the original channel measurements:

1) The original channels are reversed, such that the mobile terminal becomes the transmitting device.
2) One channel from each snapshot is kept to form a transmitter of the virtual array. This results in a virtual array with 1024 elements.
3) To reduce the correlation between adjacent channels, only 256 elements are kept using a down-sampling factor of 4.
4) Only the locations passing the chi-squared goodness of fit test for the Rayleigh fading distribution are kept.

A. Younis, W. Thompson, M. Di Renzo, C.-X. Wang, M. A. Beach, H. Haas, and P. M. Grant, "Performance of Spatial Modulation over Correlated and Uncorrelated Urban Channel Measurements", IEEE VTC-Fall, 2013. BEST PAPER AWARD.
Experimental Evaluation of SM (9/31)

![Graph showing Bit Error Ratio vs. SNR for two un-correlated channels. The graphs depict SM - Experimental, SM - Simulation, and SM - Analytical results.](image)

Fig. 1. ABER versus the SNR for SM over an uncorrelated channel. $m = 4$, $N_t = N_r = 4$. 
Experimental Evaluation of SM (10/31)

Fig. 2. ABER versus the SNR for SM over a correlated channel. $m = 4$, $N_t = N_r = 4$. 
Fig. 3. ABER versus the SNR for SM and SMX over an uncorrelated channel. $m = 4$, $N_t = N_r = 4$. 
Fig. 4. ABER versus the SNR for SM and SMX over a correlated channel. $m = 4, N_t = N_r = 4$. 
Experimental Evaluation of SM (13/31)

Fig. 5. ABER versus the SNR for SM and SMX over real measured channels. $m = 8$, $N_T = 4$. 
The binary data to be broadcast is first passed through the digital signal processing algorithm at the transmitter (DSP-Tx).

The processed data is then passed to the physical transmitter on the National Instruments (NI)-PXIe chassis (PXIe-Tx).

Each transmit antenna (‘Tx1’ and ‘Tx2’) is then activated according to the SM principle at a carrier frequency of 2.3 GHz.

The receiver then detects and processes the radio frequency (RF) signal in PXIe–Rx. Lastly, the receive side digital signal processing algorithm (DSP–Rx) recovers the original data stream.

Experimental Evaluation of SM (15/31)

Fig. 3. Binary data encoder (DSP-Tx) and decoder (DSP-Rx) algorithms for SM.
Fig. 4. Physical experimental layout: A pair of receive and a pair of transmit antennas are set 2.2 m apart from each other with a direct line of sight. Each pair of antennas is 1.5 m from the ground and there is a 10 cm spacing between the antennas in a pair corresponding to 0.77 times the wavelength at 2.3 GHz. All antennas are omnidirectional.
Digital Signal Processing for Transmission (DSP–Tx)

- The binary data is first split into information segments of appropriate size.
- The information data in each segment is then modulated using SM.
- A pilot signal used for channel estimation is then added, along with a frequency offset estimation section.
- In addition, zero-padding is performed which permits up-sampling of the data while maintaining the same signal power. The up-sampling ratio is set to four and the up-sampled data is then passed through a root raised cosine (RRC) finite impulse response (FIR) filter with 40 taps and a roll-off factor of 0.75. A large roll-off factor and a long tap-delay are necessary to ensure that the power is focused in a short time, i.e., ensure that only a single RF chain is active.
- The resulting vector is multiplied with a factor labelled ‘Tuning Signal Power’ to obtain the desired transmit power for the information sequence.
- Frames are created such that the frame length multiplied by the sampling rate is less than the coherence time of the channel which is typically ~ 7 ms for a stationary indoor environment. This ensures that all channel estimations at the receiver are valid for the frame duration.
A frame includes the frequency offset estimation sequence, the pilot and up-sampled data sequences, as shown below:

- The I16 data format is used, which is a signed 16 bit representation of an integer number.
- Each frame has at most 26100 samples.

The ‘Data section’ is formed from a series of concatenated frames.
In particular, the differences between the amplitude of the ‘Pilot and Frequency Offset’ estimation section and the amplitude of the ‘Information Data’ is clearly observable in the figure below:

- The synchronization, SNR estimation and data sections are shown
- There is approximately a 21.1 dB difference between the peak power in the synchronization section and the peak power in the SNR estimation and data sections
NI-PXIe-1075 chassis having on-board an Intel-i7 processor operating at 1.8 GHz with 4GB of RAM
Experimental Evaluation of SM (21/31)

Transmission Hardware (PXIe–Tx)

- **NI-PXIe-5450 I/Q Signal Generator**
  - 400 Mega samples (Ms)/s, 16-Bit I/Q Signal Generator
  - Dual-channel, differential I/Q signal generation
  - 512 MB of deep on-board memory
  - 16-bit resolution
  - 400 Ms/s sampling rate per channel
  - ±0.15 dB flatness to 120 MHz with digital flatness correction
  - 140 dBc/Hz phase noise density
  - −160 dBm/Hz average noise density
  - 25 ps channel-to-channel skew

- **NI-PXIe-5652 RF Signal Generator**
  - −110 dBc/Hz phase noise at 1 GHz and 10 kHz offset typical
  - 500 kHz to 6.6 GHz frequency range
  - Typically less than 2 ms frequency sweep tuning speed

- **NI-PXIe-5611 intermediate frequency (IF) to carrier RF up-converter**
The NI-PXIe-5450 I/Q signal generator is fed with the transmit vector from the binary file generated in Matlab by the encoding DSP–Tx algorithm.

In particular, the NI-PXIe-5450 I/Q signal generator performs a linear mapping of the signed 16-bit range to the output power and polarization, i.e., peak voltage amplitude is assigned to any value equal to 215 and a linear scale of the voltage amplitude down to zero.

The output from the NI-PXIe-5450 I/Q signal generator then goes to the NI-PXIe-5652 RF signal generator which is connected to the NI-PXIe-5611 frequency converter.

The NI-PXIe-5611 outputs the analogue waveform corresponding to the binary data at a carrier frequency of 2.3 GHz.

Each antenna at the transmitter and receiver contains two quarter-wave dipoles, and one half–wave dipole placed in the middle. All three dipoles are vertically polarized.

Each antenna has a peak gain of 7 dBi in the azimuth plane, with an omnidirectional radiation pattern. The 10 cm inter-antenna separation is sufficient to guarantee very low, if any, spatial correlation when broadcasting at 2.3 GHz with a 2.2 m separation between the transmitter and receiver.
Experimental Evaluation of SM (23/31)

Laboratory Setup
Experimental Evaluation of SM (24/31)

**Receiver Hardware (PXIe–Rx)**

- **NI-PXIe-1075 chassis** having on-board an Intel-i7 processor operating at 1.8 GHz with 4GB of RAM
Experimental Evaluation of SM (25/31)

Receiver Hardware (PXIe–Rx)

- NI-PXIe-5652 on-board reference clock

- NI-PXIe-5622 16-Bit Digitizer (I16)
  - 150 Ms/s real-time sampling
  - 3 to 250 MHz band in direct path mode, or 50 MHz bandwidth centered at 187.5 MHz

- NI-PXIe-5601 RF down-converter

- The receiving antennas are the same as those used for transmission

- The NIPXIe-5601 RF down-converter is used to detect the analogue RF signal from the antennas

- The signal is then sent to the NI-PXIe-5622 IF digitizer, which applies its own bandpass filter with a real flat bandwidth equal to $0.4 \times \text{SampleRate}$. The sampling rate in the experiment is 10 Ms/s which results in a real flat bandwidth of 4 MHz

- The NI-PXIe-5622 digitizer is synchronized with the NI-PXIe-5652 on-board reference clock and writes the received binary files

- The recorded binary files are then processed according to ‘DSP–Rx’
The binary files recorded by the NI-PXIe-5622 digitizer on the PXIe–Rx are converted to Matlab vectors.

In particular, a sample received vector detected by PXIe–Rx on Rx1 is as follows:
Experimental Evaluation of SM (27/31)

Digital Signal Processing for Reception (DSP–Rx)

- The Matlab vectors are then combined to form a received matrix
- The detector first finds the beginning of the transmitted sequence by using the synchronization sequence (based on an autocorrelation algorithm)
- The SNR is then calculated using the ‘SNR section’
- After the SNR for that vector has been determined, each vector is decomposed into its underlying frames
- Each frame is then down-sampled and passed through the RRC filter which completes the matched-filtering
- The frequency offset estimation, timing recovery and correction of each frame follows and are performed using state-of-the-art algorithms
- The pilot signal is then used for channel estimation
- The remaining data, along with the estimated channels, is finally used to recover an estimated binary sequence (SM maximum-likelihood demodulation)
Experimental Evaluation of SM (28/31)

Wireless Channel Characterization

CDFs of the channel coefficients

Each is defined by a Rician distribution with a unique $K$-factor

The markers denote the measurement points while the lines denote the best fit approximation.
Experimental Evaluation of SM (29/31)

The Wireline Test: RF Chain Mismatch

(b) A coaxial cable with a loss of 10 dB is connected between the transmit and receive antennas.
Experimental Evaluation of SM (30/31)

Results

- A stream of $10^5$ information bits is sent per transmission to obtain the experimental results
- The information data is put in 50, 2000 bit, frames
- The channel is estimated at the beginning and at the end of every frame resulting in 100 channel estimations per transmission
- The experiment is repeated 1000 times for every SNR point
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The Road Ahead – Open Research Challenges/Opportunities

- Appraising the Fundamental Trade-Offs of Single- vs. Multi-RF MIMO Designs
- Large-Scale Implementations: Training Overhead for CSIT/CSIR Acquisition
- From Single-User Point-to-Point to Multi-User Multi-Cell SM–MIMO Communications
- Millimeter-Wave Communications: The Need for Beamforming Gains
- Small Cell Heterogeneous Cellular Networks: Towards Interference Engineering
- Leveraging the Antenna Modulation Principle to a Larger Extent
- Open Physical-Layer Research Issues

Point-to-point SM-MIMO has been studied extensively and little room for significant steps forwards can be expected. However, some important aspects are still not completely understood:

- Transmit-diversity with single-RF base stations
- Precoding and CSIT
- Application to the uplink (co-located antennas)
- etc…

Multi-user SM-MIMO and understanding the potential of SM in cellular networks have almost been neglected so far. Here major research opportunities can be found:

- Precoding for multi-user SM-MIMO
- Application of stochastic geometry and random matrix theory to the analysis and the design of SM in HCNs
- (Low-complexity) Interference-aware SM-MIMO
- etc…
Distributed SM-MIMO for uplink applications is still almost unexplored:

- Advantages and disadvantages against state-of-the-art relaying
- End-to-end achievable diversity is unknown
- Error propagation and related low-complexity receiver design
- etc…

Energy efficiency assessment and optimization:

- The number of RF chains vs. the total number of antennas trade-off is still unclear
- Fair performance assessment and optimization against state-of-the-art
- Realistic/fair comparison with massive MIMO
- etc…

Testbed/practical implementation and measurements…
Implementation Challenges of SM-MIMO

- Antenna switching at the symbol time
- Switching loss characterization
- Reconfigurable single-RF antenna design to create unique channel signatures
- Bandwidth efficient finite-duration pulse shaping
- Large-scale antenna-array implementation and electromagnetic compatibility assessment
- Multi-carrier SM-MIMO
- Efficient channel estimation with single-RF transmitters
- Sampling time and quantization errors if orthogonal shaping filters are used
- etc…
Thank You for Your Attention

- We gratefully acknowledge the support of:
  - The European Union (ITN-GREENET project, grant 264759)
  - The Engineering and Physical Sciences Research Council (EPSRC), UK
  - The Laboratory of Signals and Systems (“Jeunes Chercheurs 2010”), France
  - The UK-China Science Bridges: R&D on (B)4G Wireless Mobile Communications
  - The Italian Inter-University Consortium for Telecommunications (CNIT), Italy
  - The European Union (ITN-CROSSFIRE project, grant 317126)
  - EADS Deutschland GmbH, Germany

M. Di Renzo  H. Haas  A. Ghrayeb
Further Readings (1/3)

Further Readings (2/3)


Further Readings (3/3)


- YouTube:
  - Spatial Modulation (http://www.youtube.com/watch?v=cPKIbxrEDho)
  - The Advantages of Spatial Modulation (http://www.youtube.com/watch?v=baKkBxf4fY)
  - The World's First Spatial Modulation Demonstration (http://www.youtube.com/watch?v=a6yUuJFUtZ4)

Further Readings – From Other Research Groups (1/3)

Further Readings – From Other Research Groups (2/3)

Further Readings – From Other Research Groups (3/3)


