Time-multiplexed single front-end multiple-input multiple-output receivers with preserved diversity gain

M. Lari\textsuperscript{1} S.A. Bassam\textsuperscript{2} A. Mohammadi\textsuperscript{1,2} F.M. Ghannouchi\textsuperscript{2}

\textsuperscript{1}Microwave and Wireless Laboratory, Electrical Engineering Department, Amirkabir University, Tehran 15914, Iran
\textsuperscript{2}Radio Lab, Schulich School of Engineering, University of Calgary, Calgary, AB, Canada T2N 1N4
E-mail: abm125@aut.ac.ir

Abstract: A design technique to realise a low-complexity multiple-input multiple-output (MIMO) receiver using a time-multiplexed single front-end receiver is introduced. The architecture uses a simple front-end receiver where it delivers radio frequency signals to baseband section. The theory behind this architecture is analysed and the diversity gain of the proposed receiver is investigated. This simple MIMO receiver can provide the same diversity gain as the diversity gain of the multiple front-end. Moreover, the experimental studies are conducted to evaluate the diversity gain of the proposed architecture. Excellent agreement between the simulated and measurement results is obtained. The proposed architecture provides a new receiver design technique with lower complexity, power consumption and cost in MIMO systems.

1 Introduction

The next generation of wireless networks demands high-speed and high-quality communication links; however, wireless channels generally provide limited bandwidth and lower quality links. The use of the diversity technique at the receiving terminal is a well-known method to mitigate the fading effects in wireless channels [1]. Single-input, multiple-output (SIMO) systems utilise multiple antennas at the receiver to achieve spatial diversity, where signals received from different antennas are passed through different fading channels. Having multiple antennas at the receiver improves the signal quality; however, it requires multiple parallel radio frequency (RF) front-ends. The baseband section processes the multiple received signals according to a spatial diversity algorithm like maximum ratio combining (MRC) [2].

The main limitation of using the multiple antenna architecture is the complexity and high cost of the hardware in the RF section. The hardware complexity and cost rise with the increase in the number of antennas. In addition, RF circuit mismatches [3] and coupling [4] grow with the increase in the number of antennas; consequently, these factors limit the use of high numbers of antennas at the transceiver.

One solution to compensate for the extra hardware cost and RF circuit imperfection is the utilisation of a single RF front-end, where a single RF path is used instead of multiple parallel RF paths. This results in an RF section that has lower complexity and cost, a simpler RF design and a compact size and power consumption.

Antenna selection [5] is a technique to reduce the cost and complexity. In an $M$ transmitting and $N$ receiving multiple-input multiple-output (MIMO) system, $L$ antennas and RF blocks, instead of $M$, is used in transmitter. On the other hand, $K$ antennas and RF blocks, instead of $N$, is used in the receiver side [6]. Antenna selection technique is investigated for diversity and spatial multiplexing [7–9]. In [10], the diversity order in an antenna selection system is analytically computed.

In [11], a novel transmitter architecture was proposed based on orthogonal transmission of multiple RF streams over a single RF front-end. The architecture in [11] focused on the transmitter side of an MIMO system. The receiver side and receiving signal processing were not a concern there, as it was assumed that the receiver was equipped with a uniform linear array with multiple RF front-ends for conventional MIMO reception. According to their approach, attention was shifted from the inputs of the diverse antenna elements to the use of different radiation patterns in the wave vector domain of the transmitting antenna array.

The idea of time-division multiplexing (TDM) technique in receivers has been used in channel sounding measurements to characterise the MIMO channel [12], in which the experimental setup used encompassed an RF switch and a single receiver, where signals from different antennas were selected, captured and digitally processed separately. Both the methods proposed in this paper and the one presented in [12] use similar receiver architectures that are based on a single RF front-end for MIMO systems; however, the purposes and hardware requirements are different. In this regard, the purpose of the switch in [12] was to selectively detect the RF signal received at each antenna and route it to the receiver. The switching speed of the switch was not critical, and it was not a function of the bandwidth of the detected signal. This contrasts with the proposed time-multiplexed single RF front-end receiving architecture, which aims at achieving maximum diversity gain for the...
MIMO receiver, requiring the capture of the signals of all the antennas for every symbol time interval of the modulated signal. This introduces constraints on the switching speed and also requires a careful alignment of the data before and after the multiplexing and de-multiplexing of the multiple signals in the MIMO receivers, as will be demonstrated in this paper. In summary, for the apparatus used in [12], in each symbol interval, the signal detected, routed and transferred from one antenna is selected from among the \( N \) antennas. This is contrary to the architecture proposed in this paper where, in each symbol interval, a time-multiplexed version of all the signals received by all the antennas is transferred to the receiver.

Using rotating antenna and parasitic elements for MIMO receiver are discussed in [13, 14]. Instead of sampling values of the electromagnetic field with antennas at discrete spatial points, they propose to oversample values of the field at the same physical location. The idea of [13] is to use parasitic elements to implement rotating antenna by single RF chain in receiver [13]. However, the complexity of implementation of rotating antenna receiver with parasitic elements is high.

The first paper that directly targets the single RF front end is [15]. The basic idea is to map multiple antenna signals into single RF chain using code division multiplexing (CDM). The authors in [15] use the orthogonal space with orthogonal codes. In this paper, we consider orthogonal space in time. We briefly discuss their approach in Section 2.2.

In this paper, we investigate the possibility of using the TDM technique to simultaneously transfer and extract the data of multiple antennas through a single RF front-end. In this regard, we first examine the possibility of using different multiplexing techniques, known as frequency division multiplexing (FDM), CDM and TDM techniques to realise a single RF front-end receiver. Then, we specifically focus on the investigation and experimental validation of the time-multiplexed single RF front-end receiver applicable in multi-antenna systems. This receiver is realised as a single RF architecture by time-multiplexing the antennas’ signals using a single-pole, multiple-throw RF switch. The received signals are de-multiplexed after RF processing according to a diversity algorithm. The simulation and measurement results show that this architecture preserves the performance of a multi-antenna system, while offering a smaller sized receiving section and providing a much lower cost and complexity for implementation.

This paper is organised as follows. First, the use of different multiplex techniques to realise a single RF front-end receiver is investigated and discussed. Then, the system model for a time-multiplexed receiver is introduced. The analytical relations and design limitations are deduced for the proposed system. This is followed by the simulation and measurement results. Finally, some concluding remarks are presented.

## 2 Multiplexing techniques for multi-antenna receiver

A conventional multi-antenna receiver with \( N \) antennas at the receiver is shown in Fig. 1. The receiver uses multiple parallel RF front-ends, where the number of RF front-ends is equal to the number of antennas. In this architecture, the baseband processing section decodes \( N \) received baseband paths to recover the signal and obtain the diversity or spatial multiplexing gain. To realise a single RF front-end path, orthogonal transmission of multiple RF streams over a single front-end must be recognised. In the following subsections, three multiplexing techniques that may be used to realise a single receiver front-end are briefly described.

### 2.1 Frequency division multiplexing (FDM)

This technique is shown in Fig. 2a. In this method, the signals of different antennas are shifted in frequency with mixing by different local oscillator and added together; therefore multiple signals of different antennas are separated in the frequency domain. After down-conversion using a single RF front-end, the frequency shifts of the multiple streams are removed in the baseband section. By further processing, the diversity gain is extracted. The method is also applicable when different signals are transmitted; for example, V-BLAST or other spatial multiplexing MIMO transmission.

A system-level performance of the scenario with two antennas at the receiving side is studied. Assuming \( s(t) \) as the passband signal with a centre frequency \( f_c \), and a bandwidth, \( \omega_0 \), the received signal of the first antenna is \( h_1 s(t) + n_1(t) \) and that of second antenna is \( h_2 s(t) + n_2(t) \), where \( n_1(t) \) and \( n_2(t) \) are also the passband noise and \( h_1, h_2 \) are the Rayleigh flat fading channel coefficients. The signal of second antenna is mixed with a low-frequency oscillator at the same frequency (\( \omega_0 \)) and filtered to remove the intermediate frequency (IF) term. The drawback of this technique is that a narrowband filter is required in the RF frequency. The bit error rate (BER) simulation result for binary phase shift keying (BPSK) modulation over 10 000 bits using two receive antennas is depicted in Fig. 3a. The difference between the ideal and simulated curves for diversities 1 and 2 is related to filter design and its quality factor.

### 2.2 Code-division multiplexing

In CDM, for example [15], the signals of different antennas must be multiplied by orthogonal codes and then added together, where a single RF front-end is used to down-convert the summed RF signals to baseband. In the baseband section, the signals are multiplied by the orthogonal codes, integrated and de-multiplexed. This technique is depicted in Fig. 2b.

In this method, the first signal is multiplied by \( c_1(t) \) code and the second signal by \( c_2(t) \) code in the symbol duration and then added together. These two codes are orthogonal, that is, \( c_1(t) \) is equal to 1 between 0 and \( T_s \), and \( c_2(t) \) is equal to 1 between 0 and \( T_s/2 \) and to −1 between \( T_s/2 \) and \( T_s \), where \( T_s \) is the symbol duration. Then, the signal \(( h_1 s(t) + n_1(t))c_1(t) + (h_2 s(t) + n_2(t))c_2(t) \) is down-converted using a single RF front-end. The baseband signal is represented as \(( h_1 \tilde{s}(t) + n_1(t))c_1(t) + (h_2 \tilde{s}(t) + n_2(t))c_2(t) \). In the baseband section, the signal is multiplied by \( c_1(t) \)
and $c_2(t)$ and separated into different paths. In each path, an integrator can remove the effect of the other signal; however, the effect of the noise of both antennas is maintained in each path.

$$h_1\tilde{s} + \tilde{n} = \frac{1}{T_s} \int_0^{T_s} [(h_1\tilde{s}(t) + \tilde{n}_1(t))c_1(t)] dt$$

$$+ \frac{1}{T_s} \int_0^{T_s} \tilde{n}_1(t) dt + 0$$

$$+ \frac{1}{T_s} \int_0^{T_s} \tilde{n}_2(t)c_1(t)c_2(t) dt$$

(1)

This is the output of the integrator in the first path where $\tilde{s}(t)$ is constant in the symbol duration $T_s$ and the codes have a unit energy. The first term of integral is $h_1\tilde{s}$ and the second part of integral is the integration of the first antenna noise. The third term is zero because two codes are orthogonal and $\tilde{s}(t)$ is invariable in the symbol duration $T_s$. The fourth term of integral is not zero although $c_1(t)$ and $c_2(t)$ are orthogonal; this is because $\tilde{n}_2(t)$ is stochastic and is not constant in the symbol duration. This term has a same power as the second part. Hence, in each sub-channel in baseband, the output noise power using CDM is twice of the noise power of traditional design.

It must be mentioned that the effect of noise is overlooked in [15]. Indeed, the noise level is increased by $10 \log(N)$ dB in each path when $N$ antennas are used and the signals are down-converted with CDM method. The simulation result for BPSK modulation over 10 000 bits using two receiving antennas is depicted in Fig. 3b. A distance of 3 dB can be observed between the conventional diversity system and this technique.

2.3 Time-division multiplexing

Fig. 2c shows a general block diagram of this architecture. This architecture uses the same number of antennas as the conventional topology; however, instead of having multiple parallel RF front-ends, a single-pole, multiple-throw RF switch along with a single RF front-end is used to down-convert the RF signals to baseband. Finally, the signals are conveyed to the baseband section using a de-multiplexer. The proposed architecture reduces the number of RF front-ends from $N$ to 1, which significantly reduces the overall cost and size of the multi-antenna receiver and also decreases the RF design mismatch.

Simulation studies were conducted using a raised cosine filter to evaluate the performance of the proposed time-multiplexed receiver. In the simulation, a BPSK signal was transmitted through a rich scatter channel. The number of receiving antennas was two, and zero forcing (ZF) receivers were used [16–18]. For each value of the signal-to-noise ratio (SNR), $10^6$ bits were transmitted.

The results of the time-multiplexed receiver using a raised cosine pulse shape filter with a roll-off factor of 0.5 for both single antenna and two antennas cases are shown in Fig. 3c. To compare the proposed architecture with a conventional receiver, the BER of the multiple front-end receivers was plotted as well. As can be seen, the diversity gain using a single time-multiplexed RF front-end was equal to the diversity gain of the multiple RF front-end. This validates the idea of using a time-multiplexed RF front-end in MIMO systems. Although the simulation shows the extraction of full diversity, the method is not restricted to diversity
systems. It can be also used when transmitter sends different data (e.g. V-BLAST [19]) from each antenna to increase data rate.

The TDM technique provides the most suitable architecture, both in performance and in hardware realisation, among the multiplexing techniques. The following sections investigate the usage of the TDM technique to realise a single front-end receiver for multi-antenna systems.

3 Theoretical analysis

3.1 Input signal

The received signal considered is a multiple quadrature amplitude modulation (M-QAM) signal with a carrier frequency $f_c$. The M-QAM modulated signal, $s_m(t)$, can be represented as

$$s_m(t) = A_{mc}u(t) \cos 2\pi f_c t - A_{ms}u(t) \sin 2\pi f_c t$$

$$m = 1, 2, \ldots, L$$

where $u(t)$ is a signal pulse shape, $f_c$ is the carrier frequency, $A_{mc}$ and $A_{ms}$ are the in-phase and quadrature signal amplitudes, $M$ is constellation size, and $L = \sqrt{M}$. The nature of the received signal is assumed to be deterministic to simplify the representation of the incoming signal in the frequency domain. The spectrum of $s_m(t)$ can, therefore, be represented as

$$S_m(f) = \frac{A_{mc}^2}{2} [U(f - f_c) + U(f + f_c)] - \frac{A_{ms}^2}{2} [-jU(f - f_c) + jU(f + f_c)]$$

where $S_m(f)$ and $U(f)$ are the Fourier transforms of $s_m(t)$ and $u(t)$, respectively.

Meanwhile, it can be easily shown that the power spectral density of M-QAM modulated signal is [20]

$$\Phi_s(f) = \frac{\sigma^2}{T} [\|U(f - f_c)\|^2 + \|U(-f - f_c)\|^2]$$

where $\sigma^2$ is the variance of the information sequence in the in-phase and quadrature paths. The spectral efficiency of the M-QAM signal is controlled by the baseband pulse shape, $u(t)$.

3.2 Switch model

According to the block diagram of the proposed architecture in Fig. 2c, the received M-QAM signals are sampled using a single-pole, $N$-throw (SPNT) RF switch. It is assumed herein that the switch takes $K$ samples from each antenna signal during the symbol period, $T$. The switch function, which obtains a sample from the first antenna, can be shown as

$$z(t) = \text{rep}_{T/K} \left[ \text{rect} \left( \frac{t}{T} \right) \right]$$

The Fourier transform of this switch waveform can be shown as

$$Z(f) = \frac{K}{T} \text{rep}_{K/T} [\delta(f)] \text{sinc}(fT)$$

Considering that the number of antennas is equal to $N$, the
switch waveform can be written as

\[
\begin{align*}
\tilde{z}\left( t \right) & = \text{rep}_{T/K} \left[ \text{rect} \left( \frac{t}{\tau} \right) + \text{rect} \left( \frac{t - (T/K)/N}{\tau} \right) \\
& + \ldots + \text{rect} \left( \frac{t - ((N-1)T/K)/N}{\tau} \right) \right] (7)
\end{align*}
\]

The sampling of different signals must use the concept of orthogonal signals in the time domain to prevent the overlapping among different samples. According to (7), in an ideal condition, the relation between the duty factor and the period is obtained as \( \tau \leq (T/K)/N \). The switch signal for \( N \) antennas in the frequency domain can be shown as

\[
\begin{align*}
Z_{\text{RF}}(f) = \frac{K}{T} \text{rep}_{T/K} [ \delta(f) ] \text{sinc} \left( f/T \right) \left\{ 1 + e^{-j2\pi fT/K/N} \right. \\
\left. + \ldots + e^{-j2\pi(N-1)T/K/N} \right\} (8)
\end{align*}
\]

### 3.3 Time sampling of received signals

According to Fig. 2c, the switch takes the sample from a particular antenna output, where the received signal arrives through a Rayleigh fading channel and is corrupted by additive white Gaussian noise (AWGN). Let us concentrate on a sample from the first antenna. This antenna signal is called \( x_1(t) \) in Fig. 2c. The signal, after passing through the filter, can be represented as

\[
x_{\text{RF}}(t) = [h_{1I}(t)A_{\text{mc}}u(t) + n_{1I}(t)] \cos(2\pi f_c t) \\
- [h_{1Q}(t)A_{\text{mc}}u(t) + n_{1Q}(t)] \sin(2\pi f_c t) (9)
\]

where \( h_{1I}(t) \) and \( h_{1Q}(t) \) are the real and imaginary parts, respectively of the channel coefficient corresponding to the first antenna, and they are assumed constant over the symbol duration, \( T \). \( n_{1I}(t) \) and \( n_{1Q}(t) \) are the real and imaginary parts, respectively of the AWGN at the output of the bandpass filter. This signal is sampled by the RF switch and is represented as

\[
x_{\text{RF}}^s(t) = [h_{1I}(t)A_{\text{mc}}u(t) + n_{1I}(t)] \cos(2\pi f_c t)z(t) \\
- [h_{1Q}(t)A_{\text{mc}}u(t) + n_{1Q}(t)] \sin(2\pi f_c t)z(t) (10)
\]

where the superscript ‘s’ represents the sampled signal.

These samples then pass through the single front-end receiver. It is assumed herein that the receiver has enough bandwidth so that RF pulse can pass through it without any distortion. The other RF samples also pass through the receiver at the different time intervals, and the samples from different antennas are orthogonal in each symbol duration. Eventually, the sample of the first antenna is delivered to the corresponding port in the baseband processing section using a low-pass filter (LPF). This port is called \( Y_1 \) in Fig. 2c. It is assumed that the RF switch and de-multiplexer are completely synchronised. The receiver is considered a simple down-convertor in this stage. If a coherent local oscillator (LO) is used at the receiver (i.e. \( x_{\text{LO}}(t) \)), the output signal, \( y_1(t) \), is expressed as

\[
y_{1i}(t) = x_{1i}(t)x_{\text{LO}}(t) (11)
\]

where the LO signals in the receiver in the in-phase and quadrature paths are \( x_{\text{LO}}(t) = \cos(2\pi f_c t) \) and \( x_{\text{LO}}(t) = \sin(2\pi f_c t) \), respectively. The signal \( y_{1i}(t) \) also encompasses the second harmonic frequency component, \( 2f_c \), which is eventually eliminated using a LPF in the baseband section. Accordingly, the output signals can be shown as

\[
y_{1I}(t) = \frac{1}{2} [h_{1I}(t)A_{\text{mc}}u(t)z(t) + n_{1I}(t)z(t)] \]
\[
y_{1Q}(t) = \frac{1}{2} [h_{1Q}(t)A_{\text{mc}}u(t)z(t) + n_{1Q}(t)z(t)] (12a)
\]

where \( y_{1I}(t) \) and \( y_{1Q}(t) \) are the in-phase and quadrature components, respectively, of the received signals. The switch effect is also considered by \( z(t) \).

For a moment, let us assume the use of a conventional receiver and a different RF front-end for each antenna. If \( z(t) = 1 \) in (12a) and (12b), these two equations are true for in-phase and quadrature components of the first antenna’s baseband received signals. In baseband, the receiver samples each antenna’s signals and performs matched filtering and MRC. If the sampler takes \( K \) samples in each symbol duration, \( T \), (12a) and (12b) are accurate for the output of the sampler when \( z(t) \) is the same as (5) and \( \tau \ll T \). Therefore (12a) and (12b) are the same for the common receiver and the time-multiplexed receiver.

### 4 Baseband processing

According to Fig. 2c, the received signals are delivered using a de-multiplexer to the baseband processing section. The LPFs are used to extract the sampled signals, which must be processed in the baseband section to extract the spatial diversity or spatial multiplexing gain. We assume that the spatial and temporal properties of the MIMO impulse response, \( H(t) \), can be separated as

\[
H(t) = u(t)H (13)
\]

where \( u(t) \) is the pulse shape.

The matched filter, \( H^*(t) \), in the MIMO baseband receiver can, therefore, be decomposed into a cascade of a space-only column matrix, \( H' \), and followed by a bank of time-only matched filters, \( u(-t) \). This is a common assumption in narrowband models [21]. It is also assumed that the channel is known on the receiver side.

In a classic multi-antenna receiver with \( N \) antennas and \( N \) RF front-ends, \( K \) samples are taken from the baseband signal of each antenna after down-conversion. The matched filter then combines the samples and improves the SNR from the samples of each antenna. Consequently, the transmitted symbol is estimated with the information of each fading channel coefficient. The same procedure can be followed in this system where there is a corresponding baseband signal for each antenna signal. We use a ZF receiver to recover the transmitted symbols [17, 18]. When the channel coefficients are known, the symbols are estimated as

\[
\hat{A}_{\text{mc}} = h_{1I}^*y_{1I} + h_{1Q}^*y_{1Q} (13a)
\]
\[
\hat{A}_{\text{mc}} = h_{1I}^*y_{1I} + h_{1Q}^*y_{1Q} (13b)
\]

where \( \hat{A}_{\text{mc}} \) and \( \hat{A}_{\text{mc}} \) are the estimated in-phase and quadrature signal amplitudes, respectively.
In this section, the simulation and experimental studies that were conducted to evaluate the performance of the proposed time-multiplexed single RF front-end receiver are described. The simulation and measurement results are presented to validate the concept of the proposed topology in practical scenarios. Experimental studies were conducted for a 16-QAM modulation scheme, where the number of receiving antennas was selected as \( N = 2 \). The 16-QAM signalling scheme was eight times oversampled and passed through a raised cosine filter with a roll-off factor of 0.3. (The final sampling rate was \( f_s = 1.6 \) MPSS (mega samples per second).) The channel model was a Rayleigh fading channel with \( f_d T_s = 0.3 \). The MRC method was used at the receiver to achieve the maximum diversity.

The objective was to show the potential of using the proposed time-multiplexed single-branch receiver, instead of conventional multi-antenna receivers. Therefore the measurement setup was planned to be as simple as possible to focus just on the functionality of the receiver, especially on the time-multiplexing and de-multiplexing concepts and to avoid the effect of other sources of distortion in the measurement setup.

The measurement setup is shown in Fig. 4. It consisted of two identical signal generators (ESG-4438C), which were used to emulate the signal received from the two antennas through uncorrelated fading channels. These two signal generators were connected through a general purpose interface bus connection and controlled with a computer.

An evaluation board from Maxim Integrated Products (MAX2830) was used as the receiver. The MAX2830 is a completely integrated solution for the implementation of RF transceivers: it contains two RF inputs, an antenna diversity switch, a low noise amplifier, a programmable voltage-gain amplifier, an RF-to-baseband down-converter and a programmable LPF. Finally, the output of the receiver board was connected to a baseband vector signal analyser (VSA), 89 600 VXI series from Agilent Inc., which was controlled with the computer through the IEEE-1394 port (FireWire).

Two signal sources transmitted the RF signal with the same carrier frequency. The faded 16-QAM modulated signals, which were generated with MATLAB software, were downloaded to each source. The outputs of the two sources were connected to the two RF inputs of the receiver board.
The received signals through the RF inputs were time-multiplexed using the integrated antenna diversity switch on the receiver board. The switch was controlled by a function generator (Agilent 33250A), where the switch speed was adjusted based on the sampling rate of the modulated signal. In our case, the sampling rate was $f_s = 1.6$ MSPS; therefore, the switch speed was set to $f_{sw} = 3.2$ MHz. The receiver board was programmed to down-convert the RF signal to a low IF signal with $f_{IF} = 3.2$ MHz. The IF signal was then captured with the VSA. Finally, an offline process was carried out in the MATLAB environment for down-conversion, de-multiplexing, MRC and LPF of the captured signal, in order to evaluate the performance of the proposed topology based on the measured BER.

A picture of the measurement setup is shown in Fig. 5. The computer was used to download the waveforms from the VSA and perform the baseband processing in the MATLAB environment.

Fig. 6 shows the measured power spectra of the received signal after time-multiplexing and RF down-conversion, but before de-multiplexing. The signal power spectra is split into three lobes: the main one was at $f_{IF} = 3.2$ MHz, and the two side lobes were at $f_{IF} \pm (f_{sw}/2) = 3.2 \pm 1.6$ MHz. This spectrum was exactly as expected from the analytical explanation in Section 3.3.

It should be noted that the proposed topology requires wider bandwidth than conventional multi-antenna receivers. This results in an extra cost; however, the overall cost is reduced, as the number of RF front-ends is decreased from $N$ in conventional multi-antenna receivers to 1 in the proposed single RF front-end receiver.

Figs. 5a and b compare the envelope of the RF signals and the extracted baseband signals related to each of the antennas after passing through the proposed receiver topology for SNR = 18 dB. The results show very good agreement between the two signals, thereby proving the functionality of the proposed architecture.

The measurement BER of the proposed time-multiplexed single RF front-end receiver is shown in Fig. 7. The measured BER results follow the analytical results of an ideal SIMO system with two antennas. The small deviation between the simulation and measurement results can be attributed to the unavoidable errors between practical measurement and theoretical simulation.

For comparison, the simulation and measurement BER results of a single-input, single-output case is also plotted in Fig. 7. As can be seen, having multiple antennas (in this case, two antennas) improved the BER performance of the communication systems due to the diversity gain. Moreover, Fig. 7 shows that the diversity gain using the proposed time-multiplexed single branch receiver was equal to the diversity gain of the multiple RF front-end. This validates the idea of using a time-multiplexed RF front-end in multi-antenna systems.

6 Conclusion

This paper discussed a novel time-multiplexed single front-end receiver suitable for MIMO systems. The architecture uses a time-multiplexed single RF front-end to down-convert the RF signal to baseband. The theory behind the proposed receiver architecture has been presented. It is shown that the time-multiplexing technique provide the most suitable implementation method to realise a single front-end receiver. Moreover, the simulation and measurement BERs for a case with two antennas, $N = 2$, was conducted to validate the performance of the proposed architecture for practical MIMO applications. The results show that using the proposed simple receiver gave the same diversity gain as conventional multiple RF front-ends. The advantages of the proposed time-multiplexed method are lower complexity, lower cost and power consumption, lower mismatch in $N$ branches and the compact size of the receiver.

7 References

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