# BEAM-PATTERN-SCANNING DYNAMIC-TIME BLOCK CODING

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### ABSTRACT

In this paper we introduce a novel transmit diversity where the space diversity in space-time block codes (STBC) is replaced by Beam Pattern Scanning (BPS). In BPS technique carefully controlled time varying phase shifts are applied to the antenna array elements mounted at the base station creating small movement in the beam pattern directed toward the desired users. In dense environments, this small movement of the antenna beam pattern creates an artificial fast fading channel leads to time diversity exploitable at the receiver. A new time dependent block coding is introduced to compensate the effect of time-varying phase shifts and ease the process of reception. Hence, we call this technique BPS/Dynamic-time block coding (BPS/DTBC). Through this merger, we achieve: 1) directionality, which supports Space Division Multiple Access (SDMA); and 2) high order of time diversity gains, which lead to a high performance. In this paper, we discuss the structure of base station antenna arrays and receiver, and we generate the probability-of-error performance simulations. These simulations show a high probability-of-error performance is achievable through the employment of BPS/DTBC scheme, compared with traditional space-time block coding.

**Index Terms** – beam pattern scanning, transmit diversity, space-time block codes, dynamic-time block codes.

## 1. INTRODUCTION

Recently, transmit diversity has been introduced as a powerful technique enhances the performance and capacity of wireless networks. Transmit diversity techniques use an array of antennas at the transmitter and/or receiver to create diversity at the receiver. Two most promising transmit diversity techniques introduced in the literature are space-time block coding (STBC) [1] [2] and beam pattern scanning (BPS) [3]-[5].

In STBC technique receiver complexity increases exponentially as a function of bandwidth efficiency, and when antenna elements at the base station are located far enough, directionality benefits are no longer available. Moreover, when large performance benefits are required, the number of antenna elements should be increased that leads to high antenna dimensions.

Beam pattern scanning (BPS) is a new transmit diversity technique capable of achieving directionality, which support Space Division Multiple Access (SDMA), and high probability-of-error performance with a low complexity and cost [3]-[5]. It uses a single antenna array at the BS to create both directionality and transmit diversity, enjoying the benefits of the two most promising antenna array applications simultaneously [4] [5] [6]. The cost of this technique is low since the complexity is mainly focused at the BS.

In this work, we introduce a novel transmit diversity technique where the space diversity in the space-time block coding is replaced by time diversity induced by BPS scheme. This technique: 1) needs a simple receiver structure to exploit the achieved diversity, 2) creates directionality that enhances the capacity of the wireless network, and 3) replaces the antenna arrays with elements located far from each other with closely located antenna elements and reduces the antenna array dimension.

In the proposed system, the data stream is applied to $N$ parallel BPS systems that share the same antenna array. Here, controlled time-varying phases applied to antenna array elements of each BPS system generate independent fades by each BPS at any time sequence within $T_s$. This makes a transmit diversity system capable of creating high quality-of-service. A new time-dependent block-code is introduced to compensate time-varying phase shifts applied at the receiver and to ease the process of reception. Hence, we call this scheme BPS dynamic-time block coding (BPS/DTBC).

Through BPS/DTBC we achieve: 1) high capacity due to Space Division Multiple Access (SDMA); and 2) high time diversity performance gain; while 3) the receiver structure is simple. This paper highlights the performance benefits achieved by BPS/DTBC in both binary phase shift keying (BPSK) and orthogonal frequency division multiplexing (OFDM) transmission.

Section 2 introduces STBC scheme and BPS technique. In Section 3, we present BPS/DTBC scheme and antenna array structure. In Section 4, we present the probability-of-error performance and discuss the complexity associated with the BPS/DTBC scheme. Section 5 concludes the paper.

## 2. STBC AND BPS TECHNIQUES

### *A. Space-time Block Codes (STBC)*

STBC is a simple transmit diversity technique capable of creating diversity at the receiver to improve the performance of communication systems [1] [2]. STBC utilizes $N$ transmit antenna separated far apart to ensure independent fades. At a given symbol period, $N$ signals are transmitted simultaneously from $N$ antennas. The signal transmitted from each antenna has a unique structure that allows the signal to be recovered at the receiver.

For simplicity in presentation, we consider an STBC with $2$ transmit antennas ($N = 2$) and one receive antenna. In addition, assuming BPSK trasnmission, the $k^{th}$ information signal entering the antenna corresponds to

$$s_k(t) = b_k \cos(\omega_o t) \cdot g_k(t)\,, \qquad (1)$$

where $b_k \in \{-1,+1\}$ is the $k^{th}$ transmitted bit, $\omega_o = 2\pi f_o$, and $f_o$ is the center frequency and $g_k(t)$ is one in $t \in [kT_s,(k+1)T_s]$ and zero otherwise. We consider $s_0$ and $s_1$ two consecutive signals generated at two consecutive times $t_0$ and $t_1 = t_0+T_s$, respectively. Via application of block codes, $s_0$ and $s_1$ are transmitted from two antennas 0 and 1 at time $t_0$, respectively, and $-s_1^*$ and $s_0^*$ (* denotes the complex conjugate operation) are transmitted from antennas 0 and 1 at time $t_1$ (See Table 1 and Fig. 1).

The channel fade is represented by:

$$\begin{aligned} h_0(t_0) &= h_0(t_0+T_S) = h_0 = \alpha_0 e^{j\theta_0} \\ h_1(t_0) &= h_1(t_0+T_S) = h_1 = \alpha_1 e^{j\theta_1} \end{aligned} \qquad (2)$$

Here, $h_0$ and $h_1$ are the channel fades for transmit antennas 1 and 2 respectively, $\alpha$ is the Rayleigh fade amplitude and $\theta$ is its phase. The channel fade amplitude is assumed Rayleigh, independent over space (antenna 0 and antenna 1), constant across two consecutive symbols (i.e., over $t_0$ and $t_1 = t_0+T_s$), and independent otherwise. Hence, the received signal can be modeled as:

$$\begin{aligned} r_0 &= r(t_0) = h_0 s_0 + h_1 s_1 + n_0 \\ r_1 &= r(t_0+T_S) = -h_0 s_1^* + h_1 s_0^* + n_1 \end{aligned} \qquad (3)$$

where $n_0$ and $n_1$ are complex Gaussian random variables representing receiver noise and interference. In the STBC receiver, the signals are combined as follows:

$$\begin{aligned} \hat{s}_0 &= h_0^* r_0 + h_1 r_1^* \\ \hat{s}_1 &= h_1^* r_0 - h_0 r_1^* \end{aligned} \qquad (4)$$

Substituting (2) and (3) into (4), we obtain

$$\begin{aligned} \hat{s}_0 &= \left(\alpha_0^2 + \alpha_1^2\right) s_0 + h_0^* n_0 + h_1 n_1^* \\ \hat{s}_1 &= \left(\alpha_0^2 + \alpha_1^2\right) s_1 - h_0 n_1^* + h_1^* n_0 \end{aligned} \qquad (5)$$

In other word, a maximum likelihood receiver leads in removal of the $s_1$ and $s_0$ dependent terms in $\hat{s}_0$ and $\hat{s}_1$, respectively. This generates a high probability-of-error performance at the receiver.

*B. Beam Pattern Scanning (BPS)*

BPS is a new transmit diversity technique utilizes an antenna array to support directionality and diversity at the same time, via applying carefully controlled time varying phase shifts to each antenna elements. This generates a small (and fast) motion of the beam pattern directed toward the desired users. As a result of the beam pattern movement, an artificial fast fading environment is created that induces time diversity exploitable by the BPS receiver.

| | *Antenna 0* | *Antenna 1* |
|---|---|---|
| Time, $t_0$ | $s_0$ | $s_1$ |
| Time, $t_0+T_S$ | $-s_1^*$ | $s_0^*$ |

Table 1: Generation of Space-Time Block Code Signals with Two Antennas

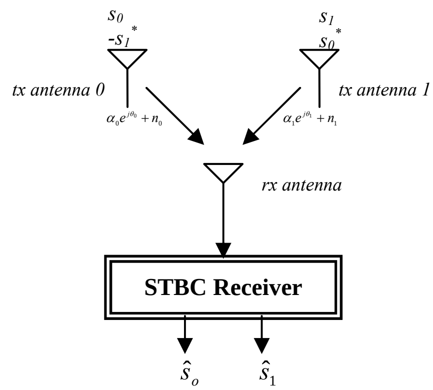



Figure 1: Space-time Block Codes System ($N$ = 2)

For ease in presentation, we will assume an $M$-element antenna array mounted horizontally at the BS (Figure 2). In order to move the antenna pattern, a time varying phase $m\theta(t)$, $m \in \{0,1,\ldots,M-1\}$ is applied to the $m^{th}$ antenna array element (Figure 2). The normalized array factor characterizing the resulting antenna pattern corresponds to

$$AF(t,\phi) = \frac{1}{M} \cdot \left[ \frac{\sin\left(\frac{M}{2}\gamma(t,\phi)\right)}{\sin\left(\frac{1}{2}\gamma(t,\phi)\right)} \right]. \qquad (6)$$

Here,

$$\gamma(t,\phi) = \frac{2\pi}{\lambda_o} d \cos\phi - \theta(t)\,, \qquad (7)$$

where $\lambda_0$ is the wavelength ($c/f_0$), $d$ is the distance between antenna elements, $\omega_o = 2\pi f_o$ where $f_0$ *is* the frequency of the transmitted signal, and time $t = 0$ refers to the first instant when the antenna array contacts a mobile user. For simplicity in presentation, it is assumed throughout that the mobile is located at angle $\phi_o \cong \pi/2$.

Through appropriate selection of $\theta(t)$, a beam pattern scanning is created. The beam pattern is *controlled* to ensure the following two criteria are satisfied:

1) Large scale fading, i.e., the mean and variance of the Rayleigh fade, is constant over symbol time $T_S$; Specifically, $\theta(t)$ controls the beam pattern to ensure that the intended receiver (mobile) remains within the beam pattern's HPBW for the entire duration $T_S$. That is:

$$\left| T_S \frac{d\phi}{dt} \right| = \kappa \cdot \beta, \quad \text{for all} \quad t \in [0, T_S], \qquad (8)$$

where $d\phi/dt$ is the rate of antenna pattern movement, $T_s \cdot d\phi/dt$ is the total antenna pattern movement in $T_s$, $\beta$ represents the antenna array HPBW, and $\kappa$, referred to as the BPS control parameter, is assigned a value $0 < \kappa < 1$. Equation (8) is solved for antenna array phase shift $\theta(t)$, leading to [4]:

$$\theta(t) = \kappa \cdot \frac{2\pi d \cdot |\sin\phi| \cdot \beta}{\lambda_o T_S} \cdot \left(t - \frac{T_S}{2}\right), \; t \in [0, T_S]. \qquad (9)$$

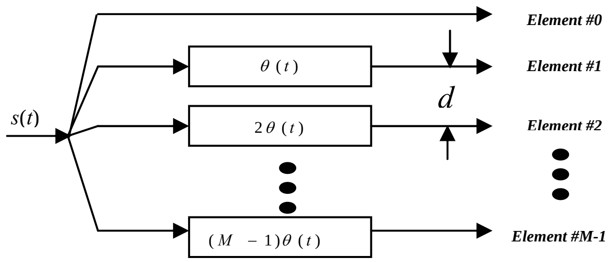



Figure 2: Antenna Array element structure

Specifically, to solve for (9) from (8), using (7), (a) we solve for $\phi$ at a fixed value of $\gamma(t,\phi) = \gamma_o$; (b) substituting this $\phi$ value into (8) and differentiating, we create a differential equation for $\theta(t)$; (c) finally, we solve the differential equation which leads directly to (9).

That is, the antenna array phases $\theta(t)$ (Fig. 2) correspond to a periodic time function to ensure a repetitive nature of the beam pattern movement (with period, $T_s$). This facilitates estimation of channel parameters over the symbol duration. The slope of the sawtooth waveform describing $\theta(t)$ is characterized by parameter $\kappa$ ($0 < \kappa < 1$, and $\kappa$ is the antenna array control parameter) and $\beta$ (the HPBW). Additionally, the slope of the sawtooth waveform determines the amount of movement in one beam pattern oscillation.

2) The BS beam pattern oscillates just enough to allow the signals received in $L$ different partitions of symbol duration $T_s$ to demonstrate independent fades. This creates an $L$-fold diversity gain at the mobile receiver. In other words, the BS antenna array sweeps the beam pattern directed at the mobile just enough to create constant large scale fading for the symbol duration $T_s$ while ensuring $L$ independent fades within each $T_s$. With beam pattern movement based on (9), which contains a control parameter $\kappa$, we set $\kappa$ to various values (in $0 < \kappa < 1$) and determine the ratio of coherence time, $T_c$ and $T_s$, i.e., $T_c/T_s$ via geometrical channel stochastic modeling and simulation (see [3]). Ratios of $T_c/T_s$, combined with Equation (9), provided us with achievable values for $L$, the diversity gain, as a function of $\kappa$, the antenna array control parameter. Assuming a medium-size city center, we found $(T_c/T_s) \approx 0.16$ whenever $\kappa \approx 0.05$ (see [3]). Using $(T_c/T_s)_{min} \approx 0.16$ and (9), we determine $L \approx 7$, i.e., a diversity of 7 fold is achievable at the receiver [3] [4].

Considering (1) as the transmitted signal over one antenna array element in Fig. 2, the total transmitted signal over all antenna elements corresponds to:

$$s_T(t) = AF(t,\phi)\cdot b_k \cdot \cos\left(\omega_o t + \frac{M-1}{2}\gamma(t,\phi)\right)\cdot g_k(t), \qquad (10)$$

where, $AF(t,\phi)$ is the array factor and $\gamma(t,\phi)$ is its phase defined in (6) and (7), respectively, in addition, $g_k(t) = 1$ for $kT_s \le t \le (k+1)T_s$. Assuming the mobile is located at $\phi_o \cong \pi/2$, in the direction of the mobile $\gamma(t,\phi) \cong \gamma(t) = -\theta(t)$ and $AF(t,\phi) \cong AF(t)$, and (10) can be approximated by,

$$s_T(t) = b_k \cdot AF(t)\cdot \cos\left(\omega_o t - \frac{M-1}{2}\theta(t)\right) g_k(t), \qquad (11)$$

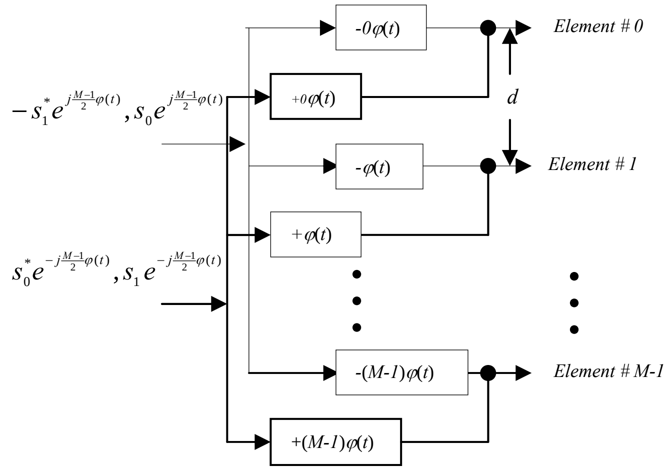



Figure 3: BPS/DTBC transmitter and antenna structure

And the received signal in $t\in[kT_s,(k+1)T_s]$ corresponds to

$$r(t) = \alpha(t)\cdot b_k \cdot AF(t)\cdot \cos\left(\omega_o t + \theta_k - \frac{M-1}{2}\theta(t) + \xi(t)\right) + n(t), \qquad (12)$$

where $\alpha(t)$ and $\xi(t)$ are the gain and phase due to the channel time varying fade, and $n(t)$ is the additive white Gaussian noise (AWGN) with variance $N_o/2$. Assuming: 1) the fade $\alpha(t)$ is constant over each duration $[lT_s,(l+1)T_s]$, and 2) the phase $\xi(t)$ is tracked and removed, the received signal in $t\in[kT_s,(k+1)T_s]$ corresponds to

$$r_l(t) = \alpha_l \cdot b_k \cdot \cos\left(\omega_o t - \frac{M-1}{2}\cdot\theta(t)\right) + n_l(t), \quad l \in \{0,1,...,L-1\} \qquad (13)$$

where $r_l(t)$ refers to the received signal over interval $[kT_s + lT_s/L,\ kT_s + (l+1)T_s/L]$ ; $n_l(t)$, $l \in \{0,1,\ldots,L-1\}$is an additive white Gaussian noise with variance $N_o/2L$; and $\alpha_l$ is a Rayleigh random variable independent for each $l$. With $\theta(t)$ selected according to (9), $\kappa \le 0.05$, the frequency offset induced by $\theta(t)$ is less than 5% and can be neglected. BPS received signal in a time sequence $l \in \{0,1,\ldots,L-1\}$ can be represented as:

$$r_l = \left(\sqrt{E_S}\,\alpha_l b_k + n_l\right)\cdot g_{k,l}(t), \qquad (14)$$

where $\sqrt{E_S}$ is the bit energy and

$$g_{k,l}(t) = \begin{cases} 1 & if \quad kT_s + lT_s/L < t < kT_s + (l+1)T_s/L \\ 0 & else \end{cases} \qquad (15)$$

Using equal gain combining (EGC) to recover the transmitted signal

$$R_C = \left(\sum_{l=0}^{L-1}\left(\alpha_l\sqrt{E_S}\,b_k + n_l\right)\right)\cdot g_k(t) \qquad (16)$$

## 3. BPS/DTBC TRANSMITTER AND RECEIVER

In BPS/DTBC, the space diversity in STBC is replaced by time diversity in BPS. Two beam patterns directed toward the desired user are generated and time varying phase shifts are controlled to move those beam patterns differently, such that, each create fades independent from the other. Figure 3 represents the proposed antenna structure. Here, one phase shift, e.g., $m\varphi(t) = m(\theta_o + \theta(t))$, $m \in \{0,1,\ldots,M-1\}$ moves the antenna beam pattern from right to left and the other one moves the antenna pattern inversely. The time-varying phase shift $\theta(t)$ is defined in (9), and $\theta_o$ is chosen in the order of $\kappa HPBW$ (e.g., *5%* of HPBW) to reduce the overlap of moving beam patterns, generate independent fades at all times and exploit full time diversity benefits at the receiver.

To compensate time-varying phase shifts in the transmitted and received signal and to ease the process of reception, the traditional block codes shown via Table 1, are replaced with dynamic-time block codes corresponding to

$$\begin{bmatrix} s_0 e^{+j\frac{M-1}{2}\varphi(t)} & s_1 e^{-j\frac{M-1}{2}\varphi(t)} \\ -s_1^* e^{+j\frac{M-1}{2}\varphi(t)} & s_0^* e^{-j\frac{M-1}{2}\varphi(t)} \end{bmatrix} \quad (17)$$

Utilizing this code structure, the received signal can be easily combined at the receiver to create a high probability-of-error performance (see Fig. 4). Considering a BPSK system, the BPS/DTBC transmitted signal for BPS system 1 and 2 at time $t_0$ with the same assumptions made for (11) corresponds to

$$\begin{aligned} s_1(t) &= \mathrm{Re}\left\{ AF^{(1)}(t)\cdot b_0 e^{j\frac{M-1}{2}\varphi(t)} \left( e^{j\left(\omega_o t - \frac{M-1}{2}\varphi(t)\right)} \right) \cdot g_0(t) \right\} \\ s_2(t) &= \mathrm{Re}\left\{ AF^{(2)}(t)\cdot b_1 e^{-j\frac{M-1}{2}\varphi(t)} \left( e^{j\left(\omega_o t + \frac{M-1}{2}\varphi(t)\right)} \right) \cdot g_0(t) \right\} \end{aligned} \quad (18)$$

Here, $AF^{(i)}(t)$, $i \in \{1,2\}$ is the array factor due the BPS system 1 and 2, where $\theta(t)$ in (7) is replaced by $\pm\varphi(t)$ for $i \in \{1,2\}$, respectively, leads to the two different antenna pattern movement. Similarly, transmitted signals from BPS systems 1 and 2 at time $t_0 + T_s$ correspond to:

$$\begin{aligned} s_1(t+T_S) &= \mathrm{Re}\left\{ AF^{(1)}(t)\cdot(-b_1) e^{j\frac{M-1}{2}\varphi(t)} \left( e^{j\left(\omega_o t - \frac{M-1}{2}\varphi(t)\right)} \right) \cdot g_1(t) \right\} \\ s_2(t+T_S) &= \mathrm{Re}\left\{ AF^{(2)}(t)\cdot b_0 e^{-j\frac{M-1}{2}\theta(t)} \left( e^{j\left(\omega_o t + \frac{M-1}{2}\varphi(t)\right)} \right) \cdot g_1(t) \right\} \end{aligned} \quad (19)$$

Equations (18) and (19) can be simplified, respectively, to:

$$\begin{aligned} s_1(t) &= \mathrm{Re}\left\{ AF^{(1)}(t)\cdot b_0 e^{j(\omega_o t)} \cdot g_0(t) \right\} \\ s_2(t) &= \mathrm{Re}\left\{ AF^{(2)}(t)\cdot b_1 e^{j(\omega_o t)} \cdot g_0(t) \right\} \end{aligned} \quad (20)$$

$$\begin{aligned} s_1(t+T_S) &= \mathrm{Re}\left\{ AF^{(1)}(t)\cdot(-b_1) e^{j(\omega_o t)} \cdot g_1(t) \right\} \\ s_2(t+T_S) &= \mathrm{Re}\left\{ AF^{(2)}(t)\cdot b_0 e^{j(\omega_o t)} \cdot g_1(t) \right\} \end{aligned} \quad (21)$$

At the receiver side, considering transmit diversity leads to $L$-fold time diversity, the received signal $[0,T_s]$ can be divided into time slots $[lT_s/L,(l+1)T_s/L]$, $l\in\{0,1,\ldots,L-1\}$, and each time slot demonstrate independent fades. The received signal at each time corresponds to:

$$\begin{aligned} r_l^{(1)}(t) &= \left(AF^{(1)}(t)\cdot\alpha_{0,l}^{(1)}\cdot b_0\cos\left(\omega_o t+\xi_{0,l}^{(1)}\right) + AF^{(2)}(t)\cdot\alpha_{0,l}^{(2)}\cdot b_1\cos\left(\omega_o t+\xi_{0,l}^{(2)}\right) + n_{0,l}(t)\right)\cdot g_{0,l}(t) \\ r_l^{(2)}(t) &= \left(AF^{(1)}(t)\cdot\alpha_{1,l}^{(1)}\cdot(-b_1)\cos\left(\omega_o t+\xi_{1,l}^{(1)}\right) + AF^{(2)}(t)\cdot\alpha_{1,l}^{(2)}\cdot b_0\cos\left(\omega_o t+\xi_{1,l}^{(2)}\right) + n_{1,l}(t)\right)\cdot g_{1,l}(t) \end{aligned} \quad (22)$$

where, $g_{k,l}(t)$ is defined in (15), $n_{k.l}(t)$ $k\in\{1,2\}$ is an additive white Gaussian noise (AWGN) independent for different time slots $l$ and $k$, $\alpha_{k,l}^{(i)}$, $i \in \{1,2\}$ is the Rayleigh fade amplitude and $\xi_{k,l}^{(i)}$ is its phase offset, in the $l \in [kT_s + lT_s/L, kT_s + (l+1)T_s/L]$, $k \in \{1,2\}$ time slot, (hereafter, this phase offset is assumed to be tracked and removed) due to BPS systems $i=1, i=2$. $\alpha_{k,l}^{(1)}$ is considered independent of $\alpha_{k,l}^{(2)}$ for any $l\in\{0,1,\ldots,L-1\}$; and, $\alpha_{k,p}^{(i)}$ is independent of $\alpha_{k,q}^{(i)}$ for all $p \neq q$ and $\alpha_{0,l}^{(i)} = \alpha_{1,l}^{(i)} = \alpha_l^{(i)}$. The receiver structure is shown in Fig. 4. The received signal, after integration, in $t \in [lT_s/L,(l+1)T_s/L]$, and assuming $AF^{(i)}(t) \approx 1$, $i \in \{1,2\}$ $AF^{(i)}(t) \approx 1, i \in \{1,2\}$ corresponds to:

$$\begin{aligned} r_{0,l} &= \frac{1}{L}\cdot\left(\alpha_l^{(1)}\cdot b_0 + \alpha_l^{(2)}\cdot b_1\right) + n_{0,l} \\ r_{2,l} &= \frac{1}{L}\cdot\left(\alpha_l^{(1)}\cdot(-b_1) + \alpha_l^{(2)}\cdot b_0\right) + n_{1,l} \end{aligned} \quad (23)$$

Signals created after block-code combining for time sequences $t\in\left[lT_S/L,(l+1)T_S/L\right]$ and $t\in\left[T_S + lT_S/L, T_S+(l+1)T_S/L\right]$, are correspond to:

$$\begin{aligned} R_{0,l} &= \frac{1}{L}\cdot\left(\left(\alpha_l^{(1)}\right)^2 + \left(\alpha_l^{(2)}\right)^2\right)\cdot b_0 + \alpha_l^{(1)} n_{0,l} + \alpha_l^{(2)} n_{1,l}^*, \\ R_{1,l} &= \frac{1}{L}\cdot\left(\left(\alpha_l^{(1)}\right)^2 + \left(\alpha_l^{(2)}\right)^2\right)\cdot b_1 - \alpha_l^{(1)} n_{1,l}^* + \alpha_l^{(2)} n_{0,l}, \end{aligned} \quad (24)$$

respectively. Equal Gain Combining (EGC) over all the time diversity components, i.e., over $l$, leads to:

$$R_0^C = \sum_{l=0}^{L-1} R_{0,l},\; R_1^C = \sum_{l=0}^{L-1} R_{1,l} \quad (25)$$

where $L$ is the order of time diversity achieved through BPS.

## 4. SIMULATION RESULTS

BPSK probability-of-error performance is simulated via BPS/DTBC, assuming a mid-size city center, i.e., $L$ = 7-fold diversity is achievable (see Fig. 5). Fig. 5(a) shows that BPS/DTBC offers 7 dB improvement at the probability-of-error of $10^{-3}$ compared to the traditional STBC. In fact, BPS/DTBC offers a probability-of-error performance very close to an AWGN channel.

Simulations are also performed for OFDM systems as shown in Fig. 5(b). It is shown that BPS/DTBC offers 17 dB, 7 dB, 4dB and 2 dB improvements comparing to the traditional OFDM, STBC/OFDM, COFDM and BPS/OFDM systems, respectively, at the probability-of-error of $10^{-3}$. The simulations clearly showed that the time and coding diversity offered by BPS/DTBC out perform other transmit diversity

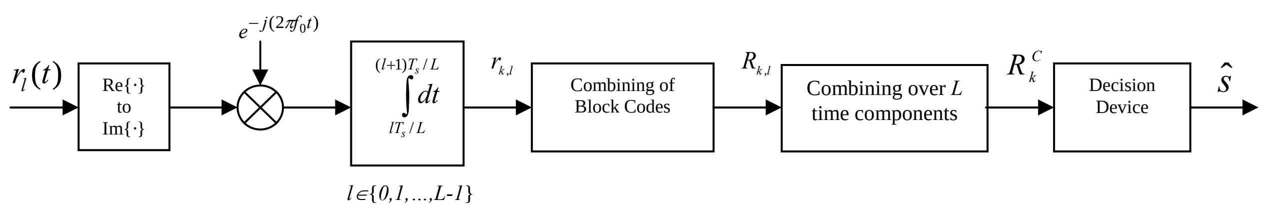
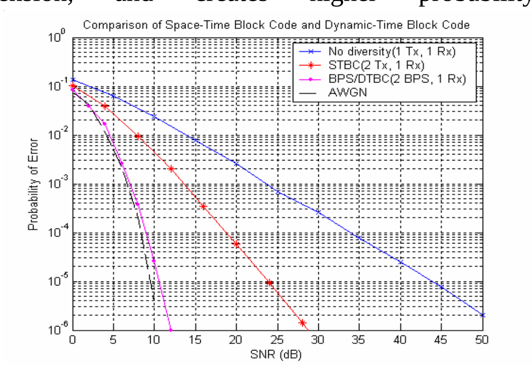



Figure 4: BPS-DTBC Receiver Design

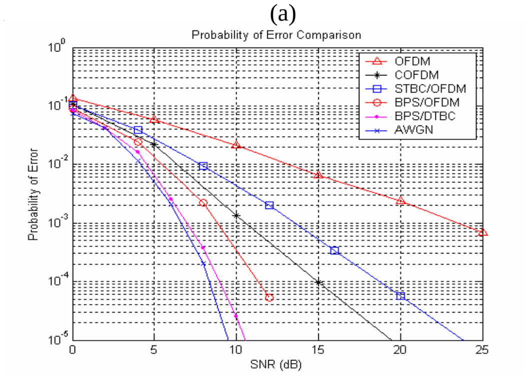
techniques such as STBC, BPS/OFDM [7] and also ½ rate forward error correction coding. This improvement shows that this scheme is a promising technique in mitigating the fading effect of a wireless communication channel.

It has to be noted that the performance achieved through BPS/DTBC is due to both transmit and coding diversities. Moreover, coding diversity benefits achieved through BPS/DTBC comes with (almost) no decrease in the overall throughput of the systems and therefore making this scheme a superior technique compared to other forward error correction coding techniques such as COFDM systems.

In the BPS/DTBC system discussed, we utilize a single antenna array at the base station, but through pre-processing we generate two sweeping antenna beam patterns. Hence, this system offers two fold diversity across BPS systems and $L$ fold (e.g., seven fold) diversity over time components, which leads to a superior probability-of-error performance. In addition, BPS/DTBC mitigates the space requirement in the traditional STBC: In traditional STBC, antennas are required to be located far enough (e.g., in the order of $5\lambda$) to ensure the signal transmitted by each antenna experiences independent fades. However, in BPS/DTBC, antenna arrays are separated by $\lambda/2$ to create directionality. Therefore, for an $M = 6$ element antenna arrays the dimension required is only $2.5\lambda$ which is less than the dimension needed for STBC. Hence, BPS/DTBC creates that amount of performance with a smaller antenna array dimensions. Finally, since antenna arrays in this system create directionality, it enhances wireless network capacity via Spatial Division Multiple Access (SDMA). Hence, BPS/DTBC uses an antenna array with a smaller dimension, and creates higher probability-of-error performance and network capacity. The cost of this system is low since the complexity is mainly focused at the bases station; moreover, the BPS-DTBC receiver complexity is relatively low because the diversity components enter the receiver serially in time.




(a)




(b)

Figure 5: a) BPSK system, b) OFDM system

## 5. CONCLUSIONS

The innovative merger of BPS and STBC introduced and its probability-of-error was studied for both BPSK and OFDM systems. We called this scheme BPS/DTBC that creates time diversity through BPS and coding diversity through dynamic time block codes highly enhances the probability-of-error performance of the system. BPS/DTBC achieves this improvement via: 1) antennas with a smaller physical dimension compared to the traditional STBC, 2) antenna arrays that support SDMA, leads to higher network capacity, and 3) lower cost. This makes BPS/DTBC a possible strong transmit diversity scheme for future wireless communications systems.

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