A Spatial Multiplexing Scheme of Overlapping Coverage Area for Single Frequency Networks

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Abstract

The "artificial multipath" in the overlapping coverage area of the single frequency network (SFN) increases the signal power. However, the generated self-interference due to SFN overlapping will also worsen the signal-to-noise ratio (SNR) condition. This work proves that the SFN overlapping coverage area can achieve the maximal transmit diversity gain in the form of orthogonal spatial multiplexing. Based on this, a parallel Alamouti orthogonal spatial multiplexing scheme is reconstruct to promote the spatial freedom, obtain the transmit diversity gain and overcome the deterioration of signal quality in the overlapping coverage area. Simulation results show that when the number of the transmitting antennas is two and the number of the receiving antenna is one, the introduced parallel Alamouti-SFN scheme can achieve 9.5dB SNR gains when BER is 10^{-3} compared with the traditional SFN scheme. It is verified that the orthogonal spatial multiplexing scheme can effectively utilizes the self-interference of overlapping coverage areas to improve signal quality.

Keywords: single frequency network, overlapping coverage area, parallel Alamouti spatial multiplexing, multiple input, multiple output technology

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1. Introduction

The single frequency network can improve the coverage uniformity and reduce the transmitter cost by reasonably setting the position of each station, the power of the transmitting signal and the height of the transmitting and receiving antennas [1]. With the rapid development of high-speed train (HST), the application of SFN in high-speed mobile communication network has received more and more attention [2]. To cope with the increasingly fierce spectrum competition, Europe announced the second-generation digital terrestrial TV standard DVB T2 in 2009, China initiated the research and development plan of DTMB-A and was included in the ITU-R proposal on July 8, 2015. The new standards has effectively improved the performance of SFN system by supporting advanced transmission diversity technology [3].

On 1 August 2016, Japanese company NHK started to broadcast 8K ultra-high definition TV (UHDTV) via satellite, and established two SFN experimental stations using space-time coding technology to evaluate the performance of the next generation ground transmission network [4].

There are serious multipath interference problems in the networking mode of SFN [5]. In SFN system, except for the "natural multipath" caused by the reflection and refraction of electromagnetic waves, a number of "artificial multipath" will be existing due to the same signals sent by different base stations with same frequency reaching the receiver through different propagation paths [6].

[7] verified the overlapping defects of SFN through experimental tests. When the delay spread of the two signals does not exceed the system guard interval size, the received carrier-to-noise ratio (C/N) required for error-free reception is approximately doubled. So a higher C/N is required to resist the interference of the 0dB echo on the receiving signal, which also offsets the increase of receiving signal power accordingly.

In this paper, the signal reception pattern of the SFN overlapping coverage area is equivalent to the scene of multiple transmitting antenna system. Furthermore, an Alamouti orthogonal spatial multiplexing scheme is proposed to promote the spatial freedom, obtain the transmit diversity gain. Utilizing parallel Alamouti-SFN scheme and the corresponding Space-time decoding, the deterioration of signal quality in the overlapping coverage area will be overcame, meanwhile normal reception performance of non-overlapping coverage areas can be maintained. The orthogonal coded signal format design can also greatly simplify the implementation complexity of the receiver. The experimental results verified the orthogonal spatial multiplexing can effectively utilizes the self-interference of overlapping coverage areas to improve signal quality.

This paper is organised as follows:

In Section 2, a SFN system composed of two transmitting stations is used as an example to model the channel and analyze the coverage characteristics of SFN overlapping area.

In Section 3, SFN overlapping coverage area performance of Alamouti spatial multiplexing is discussed.

Parallel Alamouti-SFN networking solution is discussed in Section 4.

Section 5 simulates and compares the error performance of the parallel Alamouti-SFN scheme and traditional SFN scheme in overlapping coverage area and non-overlapping coverage area.

2. SFN System

2.1 SFN channel model

In the SFN system, transmitting stations at different locations can use the same frequency to send the same information (Figure 1).

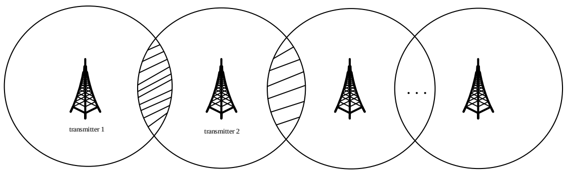

Figure 1. Single frequency network overlapping coverage area model

As shown in Figure 1, the double base stations scenario utilizing relay-style SFN considered in this paper is the coverage of long and narrow areas such as highways.

Assume the transmitting signal at time t is (t) , the receiving signal $r(t)$ can be obtained by convolving with the channel impulse response $h_i(\tau,t)$:

$$r_i(t) = \sqrt{p_{ri}}\sum_{m=0}^{M_i-1}\rho_m e^{j\theta_m}s(t-\tau_m)+w(t) \quad i=1,2 \tag{1}$$

$$h_i(\tau,t) = \sum_{m=0}^{M_i-1}\rho_m e^{j\theta_m}\delta(t-\tau_m) \tag{2}$$

where

- M_i refers to the multipath number of the transmitter to the receiver,
- ρ_m, θ_m, τ_m represent the amplitude, phase, and delay of the m-th path respectively,
- $w(t)$ is the additive white Gaussian noise superimposed on the signal;
- $p_{ri} = a * \frac{p_{ti}}{l_i^{\alpha}}$ represents the receiving signal power under the influence of large scale fading, where l_i is the path distance , p_{ti} is the power of the transmitting signal and a is the scaling factor of large scale fading [8].

Consider the effects of natural multipath and “artificial multipath”, the receiving signal in overlapping coverage area can be written as [9]:

$$r(t) = \sqrt{p_{r1}}\sum_{m=0}^{M_1-1}\rho_m e^{j\theta_m}s(t-\tau_m)+ \sqrt{p_{r2}}\sum_{m=0}^{M_2-1}\rho_m e^{j\theta_m}s(t-\tau_m)+w(t) \tag{3}$$

Assume that the distance between the two transmitting stations is $2d$. The line connecting the two stations is the horizontal axis, and the vertical bisector is the vertical axis.

The rectangular coordinate system as shown in Figure 2 is established.

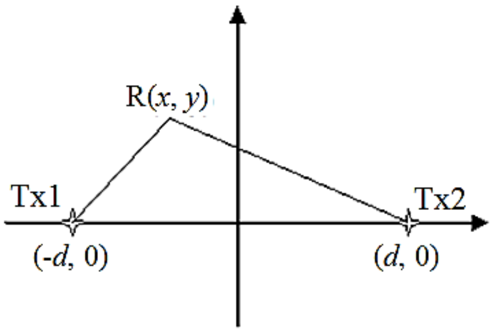

Figure 2. Single frequency network overlapping coverage area model

Take the signal of transmitter 1 (Tx1) as reference, the power of the receiving signal of transmitter 2 (Tx2) relative to Tx1 can be replaced as follows:

$$\gamma = \frac{p_{t2}*\left(\sqrt{(x+d)^2+y^2}\right)^2}{p_{t1}*\left(\sqrt{(x-d)^2+y^2}\right)^2} \tag{4}$$

We assume that the transmission delay and the path delay of the Tx2 relative to the Tx1 are τ_t and τ_p respectively. Let $\tau_t = k_t\,T_S, \tau_p = k_p\,T_S, \tau_m = k_m\,T_S$, where T_S is the sample interval. Omitting T_S, (3) can be rewritten as [10]:

$$r_n = r_1(n)+r_2(n)+w(n) = \sum_{m=0}^{M_1-1}\rho_m e^{j\theta_m}s(n-k_m)+ \sqrt{\frac{p_{t2}*\left(\sqrt{(x+d)^2+y^2}\right)^2}{p_{t1}*\left(\sqrt{(x-d)^2+y^2}\right)^2}}\sum_{m=0}^{M_2-1}\rho_m e^{j\theta_m}s\left(n-k_m-k_t-k_p\right)+ w(n) \tag{5}$$

where $w(n)$ represents the thermal noise superimposed on the receiving antenna.

2.2 Main defect of SFN overlapping coverage area

In SFN system, the receiving signal is the mixture of transmitting signals with different delay. The receiving signal power in SFN overlapping coverage area can be described as:

$$p_r = p_{r1+}p_{r2} = a*\frac{p_{t1}}{\left(\sqrt{(x+d)^2+y^2}\right)^2}+a*\frac{p_{t2}}{\left(\sqrt{(x-d)^2+y^2}\right)^2} \tag{6}$$

The power gain of SFN overlapping coverage area can be obtained as follows [11]:

$$G = \frac{p_{r1+}p_{r2}}{p_{r1}} = 1+\frac{p_{t2}*\left(\sqrt{(x+d)^2+y^2}\right)^2}{p_{t1}*\left(\sqrt{(x-d)^2+y^2}\right)^2} \tag{7}$$

When the receiver is on the vertical bisector line of two transmitting stations, the coordinate of the receiver is $(0, y)$, the power gain is:

$$G = 1+\frac{p_{t2}}{p_{t1}} > 1 \tag{8}$$

Furthermore, the receiving signal power ratio of the receiver is equal to the transmitting power ratio of the two transmitting stations, i.e.:

$$\frac{p_{r2}}{p_{r1}} = \frac{p_{t2}}{p_{t1}} \tag{9}$$

The receiving signal power of overlapping coverage area in SFN has gain relative to the area of a single transmitting station. From the perspective of power, the transmission condition of SFN is favorable for signal reception.

According to (5), signals from differernt transmitting stations arrive at the receiver with different amplitudes and phases. Since the signal strengths of the two transmitting stations are similar, some "artificial multipath" of 0dB or approximately 0dB is generated under directional antenna conditions.

In [7], laboratory tests have found that when the signals transmitted by two base stations in SFN arrive at the receiver in the same size, the power of the rec eiving signal is doubled. When the delay spread of the two signals does not exceed the system guard interval, the the received C/N value required for error-free reception is approximately doubled. So, a higher C/N is required to resist the interference of the 0 dB echo on the receiving signal, which cancels out the increase of the power of the receiving signal accordingly. Under the same condition , the extreme case of the 0 dB echo can be tolerated in the SFN system, it is difficult to bring the transmit diversity due to the increase of the required C/N threshold. Therefore, it exists a field that receiving power is deteriorating [12].

Next, an improved Alamouti spatial diversity technology is proposed to reconstruct multi-station SFN system, which reduces the possibility that the signal amplitude suffers from deep fading and improves the error characteristics.

3. SFN overlapping coverage area performance of Alamouti spatial multiplexing

3.1 Diversity gain

The space diversity technique can obtain the diversity gain by encoding the transmitting signal in both time and space, and reduce the error probability of decoding [13].

Assume that the total number of space-time symbols transmitted is N and $N = N_T * T$, where N_T is the number of transmitting antennas and T is the time required to transmit these symbols. In Rayleigh channel, the receiving baseband sequence can be expressed as [14]:

$$\mathbf{R} = \sqrt{SNR}\mathbf{H}\mathbf{S}^{\mathbf{T}} + \mathbf{W} \tag{10}$$

where

$\mathbf{S} = \begin{Bmatrix} s_1^{(1)} & \cdots & s_1^{(T)} \\ \vdots & \ddots & \vdots \\ s_{N_T}^{(1)} & \cdots & s_{N_T}^{(T)} \end{Bmatrix}$ is the transmitting signal matrix;

$\mathbf{H} = \begin{Bmatrix} h_{11} & \cdots & h_{1N_T} \\ \vdots & \ddots & \vdots \\ h_{N_R1} & \cdots & h_{N_RN_T} \end{Bmatrix}$ is the channel matrix, where h_{ij} represents the channel fading from the i-th transmitting antenna to the j-th receiving antenna;

$\mathbf{w} = \begin{Bmatrix} w_1^{(1)} & \cdots & w_1^{(T)} \\ \vdots & \ddots & \vdots \\ w_{N_T}^{(1)} & \cdots & w_{N_T}^{(T)} \end{Bmatrix}$is the noise interference.

$s_j^{(t)}, w_i^{(t)}, h_{ij}$ are cyclic symmetric complex Gaussian random variables with a mean of 0 and a variance of 1.

$SNR = \frac{E_S}{N_0 N_T}$ is the signal to noise ratio on the average transmitting antenna in each symbol cycle.

If the channel state information (CSI) is obtained at the receiver, the codeword matrix $\hat{\mathbf{S}}$ can be decoded by the maximum likelihood (ML) decoding algorithm.

When (11) is established, bits errors can occuring in decoding [15].

$$-\sqrt{\frac{E_S}{N_0 N_T}}\sum_{t=1}^{T}\sum_{i=1}^{N_R} 2\mathrm{Re}\left\{\sum_{j=1}^{N_T} h_{ij}\left(s_j^{(t)} - \hat{s}_j^{(t)}\right)\cdot\left(w_i^{(t)}\right)^*\right\} \geq \sum_{t=1}^{T}\sum_{i=1}^{N_R}\left|\sqrt{\frac{E_S}{N_0 N_T}}\sum_{j=1}^{N_T} h_{ij}\left(s_j^{(t)} - \hat{s}_j^{(t)}\right)\right|^2 \tag{11}$$

The right side of (11) can be rewritten as

$\frac{E_S}{N_0 N_T}\left\|\mathbf{H}(\mathbf{S} - \hat{\mathbf{S}})\right\|^2$.

The left side of (11) can then be written as:

$$-\sqrt{\frac{E_S}{N_0 N_T}}\sum_{t=1}^{T}\sum_{i=1}^{N_R} 2\mathrm{Re}\left\{\sum_{j=1}^{N_T} h_{ij}\left(s_j^{(t)} - \hat{s}_j^{(t)}\right)\cdot\left(w_i^{(t)}\right)^*\right\} = -\sqrt{\frac{E_S}{N_0 N_T}}\sum_{i=1}^{N_R} \boldsymbol{e}_i 2\mathrm{Re}\left\{\mathbf{H}\left(\mathbf{S} - \hat{\mathbf{S}}\right)\cdot\mathbf{W}^{\mathrm{H}}\right\}\boldsymbol{e}_i^{\mathrm{T}} = 2\sqrt{\frac{E_S}{N_0 N_T}}\sum_{i=1}^{N_R} \mathrm{Im}\left\{\boldsymbol{h}_i\left(\mathbf{S} - \hat{\mathbf{S}}\right)\right\}\cdot\mathrm{Im}\left\{\boldsymbol{w}_i^{\mathrm{H}}\right\} - \mathrm{Re}\left\{\boldsymbol{h}_i(\mathbf{S} - \hat{\mathbf{S}})\right\}\cdot\mathrm{Re}\left\{\boldsymbol{w}_i^{\mathrm{H}}\right\} \tag{12}$$

where

$\boldsymbol{e}_i = [0_1 0_2 \cdots \boldsymbol{I}_i \cdots 0_{N_R}]$,

$\sum_{i=1}^{N_R} 2\boldsymbol{e}_i\mathrm{Re}\left\{\mathbf{H}\left(\mathbf{S} - \hat{\mathbf{S}}\right)\cdot\mathbf{W}^{\mathrm{H}}\right\}\boldsymbol{e}_i^{\mathrm{T}}$ represents the sum of the elements on the main diagonal;

the vector $\boldsymbol{h}_i$ represents the i-row of $\mathbf{H}$,

$\boldsymbol{w}_i^{\mathrm{H}}$ represents the i -column of $\mathbf{W}^{\mathrm{H}}$.

When (12) is subject to the gaussian distribution of zero mean, and the variance σ^2 can be rewritten as:

$$\sigma^2 = \left(2\sqrt{\frac{E_S}{N_0 N_T}}\right)^2\left\{\left|\sum_{i=1}^{N_R}\mathrm{Im}\left\{\boldsymbol{h}_i\left(\mathbf{S} - \hat{\mathbf{S}}\right)\right\}\right|^2 + \left|\sum_{i=1}^{N_R}\mathrm{Re}\left\{\boldsymbol{h}_i\left(\mathbf{S} - \hat{\mathbf{S}}\right)\right\}\right|^2\right\}\times\frac{1}{2} = 2\frac{E_S}{N_0 N_T}\left\|\mathbf{H}(\mathbf{S} - \hat{\mathbf{S}})\right\|^2 \tag{13}$$

Normalize (13) into a standard normal distribution as follows:

$$\Gamma \triangleq \frac{left}{\sqrt{\sigma^2}} \geq \frac{\frac{E_S}{N_0 N_T}\left\|\mathbf{H}(\mathbf{S}-\hat{\mathbf{S}})\right\|^2}{\sqrt{2\frac{E_S}{N_0 N_T}\left\|\mathbf{H}(\mathbf{S}-\hat{\mathbf{S}})\right\|^2}} = \sqrt{\frac{E_S}{2N_0 N_T}}\left\|\mathbf{H}(\mathbf{S} - \hat{\mathbf{S}})\right\| \tag{14}$$

The paired error probability based on $\mathbf{H}$ can be obtained:

$$\mathrm{P}\left(\mathbf{S}, \hat{\mathbf{S}}\middle|\mathbf{H}\right) = \mathrm{Q}\left(\sqrt{\frac{E_S}{2N_0 N_T}}\left\|\mathbf{H}(\mathbf{S} - \hat{\mathbf{S}})\right\|\right) \tag{15}$$

where Q(x) function satisfies [16]:

$$\mathrm{Q}(\mathrm{x}) \leq \frac{1}{2}\exp\left(\frac{-x^2}{2}\right) \tag{16}$$

The upper limit of paired error probability can be rewritten as follows:

$$\mathrm{P}\left(\mathbf{S}, \hat{\mathbf{S}}\middle|\mathbf{H}\right) \leq \frac{1}{2}\exp\left(-\frac{E_S}{4N_0 N_T}\left\|\mathbf{H}(\mathbf{S} - \hat{\mathbf{S}})\right\|^2\right) \tag{17}$$

To minimize the upper limit of paird error probability, maximize the value of $\left\|\mathbf{H}(\mathbf{S} - \hat{\mathbf{S}})\right\|^2$.

$$\left\|\mathbf{H}\left(\mathbf{S} - \hat{\mathbf{S}}\right)\right\|^2 = \mathrm{Tr}\left(\mathbf{H}\left(\mathbf{S} - \hat{\mathbf{S}}\right)\mathbf{H}\left(\mathbf{S} - \hat{\mathbf{S}}\right)^{\mathrm{H}}\mathbf{H}^{\mathrm{H}}\right) = \sum_{i=1}^{N_R}\boldsymbol{h}_i(\mathbf{S} - \hat{\mathbf{S}})(\mathbf{S} - \hat{\mathbf{S}})^{\mathrm{H}}\,\boldsymbol{h}_i^{\mathrm{H}} \tag{18}$$

Let $\mathbf{A}\left(\mathbf{S} - \hat{\mathbf{S}}\right) = \left(\mathbf{S} - \hat{\mathbf{S}}\right)\left(\mathbf{S} - \hat{\mathbf{S}}\right)^{\mathrm{H}}$. Due to $\mathbf{A}$ is equal to its conjugate transposed matrix, SVD decomposition on matrix $\mathbf{A}\left(\mathbf{S} - \hat{\mathbf{S}}\right)$ can be operated as:

$$\mathbf{A}\left(\mathbf{S} - \hat{\mathbf{S}}\right) = \mathbf{V}\Delta\mathbf{V}^{\mathrm{H}} \tag{19}$$

where $\mathbf{V}$ is the unitary matrix, element of Δ is the eigenvalue of matrix $\mathbf{A}$[17].

Meanwhile, (18) can be simplified as:

$$\sum_{i=1}^{N_R} \boldsymbol{h}_i(\mathbf{S}-\hat{\mathbf{S}})(\mathbf{S}-\hat{\mathbf{S}})^{\mathrm{H}} \boldsymbol{h}_i^{\mathrm{H}} = \sum_{i=1}^{N_R} \boldsymbol{h}_i \mathbf{V}\Delta\mathbf{V}^{\mathrm{H}}\boldsymbol{h}_i^{\mathrm{H}} = \sum_{i=1}^{N_R}\sum_{j=1}^{N_T} \lambda_i \left|\boldsymbol{h}_i\mathbf{v}_j\right|^2 \quad (20)$$

Let $\beta_{ij} = |\boldsymbol{h}_i\mathbf{v}_j|$, (17) can be simplified as:

$$\mathrm{P}(\mathbf{S},\hat{\mathbf{S}}|\mathbf{H}) \le \frac{1}{2}\exp(-\frac{E_S}{4N_0N_T}\sum_{i=1}^{N_R}\sum_{j=1}^{N_T}\lambda_j\left|\beta_{ij}\right|^2) \quad (21)$$

When $\beta_{ij} = |\boldsymbol{h}_i\mathbf{v}_j|$, the distribution of β_{ij} is mainly determined by $\boldsymbol{h}_i$.

Probability density function of β_{ij} in Rayleigh fading channel can be described as follows:

$$f(|\beta_{ij}|) = 2|\beta_{ij}|\exp(-|\beta_{ij}|^2) \quad (22)$$

The paired error probability based on $\mathbf{H}$ can be rewritten as:

$$\mathrm{P}(\mathbf{S},\hat{\mathbf{S}}|\mathbf{H}) \le \left(\sum_{j=1}^{N_T}\frac{1}{1+\frac{E_S}{4N_0}\lambda_j}\right)^{N_R} \quad (23)$$

When $\frac{E_S}{N_0}$ is large enough, (23) in dB can be expressed as:

$$\lg\left(\mathrm{P}(\mathbf{S},\hat{\mathbf{S}}|\mathbf{H})\right) \le -rN_R\lg\left(\left(\prod_{j=1}^{r}\lambda_j\right)^{1/r}\left(\frac{E_S}{4N_0}\right)\right) \quad (24)$$

where r is the rank of $\mathbf{A}$.

The upper bound of $\mathrm{P}(\mathbf{S},\hat{\mathbf{S}}|\mathbf{H})$ is mainly determined by the size of rN_R, so rN_R is equivalent to the diversity gain.

3.2 Diversity effect of SFN

In traditional SFN design, two transmitting stations send the same data stream [18]. Assume that the length of transmitting sequence is L, the transmitting signal matrix can be described as follows:

$$\mathbf{S} = \begin{pmatrix} s_1 & \cdots & s_L \\ s_1 & \cdots & s_L \end{pmatrix} \quad (25)$$

The receiver performs ML decoding on the receiving signal, and the result is denoted as $\hat{\mathbf{S}}$.

Then, the code word difference matrix $\mathbf{B}(\mathbf{S},\hat{\mathbf{S}})$ can be written as:

$$\mathbf{B}(\mathbf{S},\hat{\mathbf{S}}) = \begin{pmatrix} s_1-\hat{s}_1 & \cdots & s_L-\hat{s}_L \\ s_1-\hat{s}_1 & \cdots & s_L-\hat{s}_L \end{pmatrix} \quad (26)$$

Code word distance matrix can be computed as follows:

$$\mathbf{A}(\mathbf{S},\hat{\mathbf{S}}) = \mathbf{B}(\mathbf{S},\hat{\mathbf{S}})\mathbf{B}^{\mathrm{H}}(\mathbf{S},\hat{\mathbf{S}})$$
$$\begin{pmatrix} |s_1-\hat{s}_1|^2+\cdots+|s_L-\hat{s}_L|^2 & |s_1-\hat{s}_1|^2+\cdots+|s_L-\hat{s}_L|^2 \\ |s_1-\hat{s}_1|^2+\cdots+|s_L-\hat{s}_L|^2 & |s_1-\hat{s}_1|^2+\cdots+|s_L-\hat{s}_L|^2 \end{pmatrix} \quad (27)$$

Performe elementary row transformation on matrix $\mathbf{A}(\mathbf{S},\hat{\mathbf{S}})$:

$$\mathbf{A}(\mathbf{S},\hat{\mathbf{S}}) = \mathbf{B}(\mathbf{S},\hat{\mathbf{S}})\mathbf{B}^{\mathrm{H}}(\mathbf{S},\hat{\mathbf{S}})$$
$$= \begin{pmatrix} |s_1-\hat{s}_1|^2+\cdots+|s_L-\hat{s}_L|^2 & |s_1-\hat{s}_1|^2+\cdots+|s_L-\hat{s}_L|^2 \\ 0 & 0 \end{pmatrix} \quad (28)$$

This work proposed a parallel Alamouti spatial diversity scheme for SFN, in which the same diversity gain as the maximum ratio combined reception (MRC) and lower reception complexity can be obtained without extending the transmission bandwidth [19]. In this SFN design, after parallel Aiamouti spatial multiplexing the transmitting signal matrix can be described as follows:

$$\mathbf{S} = \begin{pmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{pmatrix} \quad (29)$$

The code word difference matrix $\mathbf{B}(\mathbf{S},\hat{\mathbf{S}})$ can be written as follows:

$$\mathbf{B}(\mathbf{S},\hat{\mathbf{S}}) = \begin{pmatrix} s_1-\hat{s}_1 & s_2-\hat{s}_2 \\ -s_2^*+\hat{s}_2^* & s_1^*-\hat{s}_1^* \end{pmatrix} \quad (30)$$

Due to the orthogonality of the coding matrix, $\mathbf{B}(\mathbf{S},\hat{\mathbf{S}})$ is also orthogonal. The code word distance matrix can be computed as follows:

$$\mathbf{A}(\mathbf{S},\hat{\mathbf{S}}) = \mathbf{B}(\mathbf{S},\hat{\mathbf{S}})\mathbf{B}^{\mathrm{H}}(\mathbf{S},\hat{\mathbf{S}})$$
$$= \begin{pmatrix} |s_1-\hat{s}_1|^2+|s_2-\hat{s}_2|^2 & 0 \\ 0 & |-s_2^*+\hat{s}_2^*|^2+\cdots+|s_1^*-\hat{s}_1^*|^2 \end{pmatrix} \quad (31)$$

In the traditional SFN design, the rank of matrix $\mathbf{A}$ is $r=1$. For one receiving antenna, the diversity gain $rN_R=1$. While designing in Alamouti spatial multiplexing, the rank of matrix $\mathbf{A}$ is $r=2$. For one receiving antenna, the maximal transmit diversity gain $rN_R=2$.

4. SFN networking based on Alamouti spatial multiplexing

Parallel Alamouti spatial multiplexing system implementation model is shown in Figure 3.

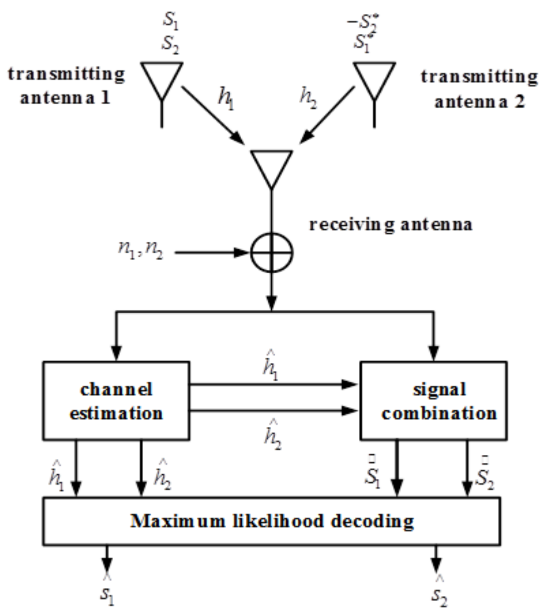

Figure 3. The parallel Alamouti spatial multiplexing system

For the linearly distributed SFN system, when the transmission stations are far apart, it can be considered that the same receiver can only receive signals from two transmitting stations at most [20]. Such SFN is distributed in narrow areas such as urban roads and streets. After the introduction of parallel Alamouti spatial multiplexing, the coded transmitter and the uncoded transmitter are alternated, and the overall structure is shown in Figure 4.

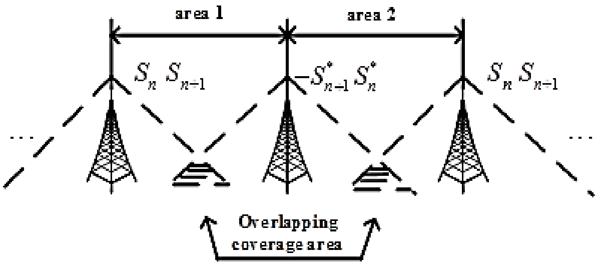

Figure 4. The linearly distributed SFN coding model

After two adjacent OFDM symbols pass through the MISO channel, the receiving signal in (5) can be expressed as:

$$r_1^n = h_1 s_n - h_2 s_{n+1}^* + w_1^n \tag{32}$$

$$r_1^{n+1} = h_1 s_{n+1} + h_2 s_n^* + w_1^{n+1} \tag{33}$$

where w_1^n and w_1^{n+1} are the additive noise of two adjacent OFDM symbol periodic channels.

The signals received in the n and $n+1$ symbol cycles can be written in matrix form:

$$\begin{bmatrix} r_1^n \\ (r_1^{n+1})^* \end{bmatrix} = \begin{bmatrix} h_1 & -h_2 \\ h_2^* & h_1^* \end{bmatrix} \begin{bmatrix} s_n \\ s_{n+1}^* \end{bmatrix} + \begin{bmatrix} w_1^n \\ (w_1^{n+1})^* \end{bmatrix} \tag{34}$$

Using the ML decoding method, the codeword matrix can be obtained as follows:

$$\hat{s}_n = \arg \min_{\hat{s}_n \in \mathbf{C}} (|h_1|^2 + |h_2|^2 - 1)|\hat{s}_n|^2 + d^2(\tilde{s}_n, \hat{s}_n) \tag{35}$$

$$\hat{s}_{n+1} = \arg \min_{\hat{s}_{n+1} \in \mathbf{C}} (|h_1|^2 + |h_2|^2 - 1)|\hat{s}_{n+1}|^2 + d^2(\tilde{s}_{n+1}, \hat{s}_{n+1}) \tag{36}$$

where $\mathbf{C}$ represents the set of valid code words for all constellation maps, and $(|h_1|^2 + |h_2|^2)$ reflects that the spatial diversity gain is 2.

When one of the two antennas can't work, for example, assume that the receiver can only receive the signal transmitted by Tx1:

$$r_1^n = h_1 s_n + w_1^n \tag{37}$$

$$r_1^{n+1} = h_1 s_{n+1} + w_1^{n+1} \tag{38}$$

$$\begin{bmatrix} \tilde{s}_n \\ \tilde{s}_{n+1} \end{bmatrix} = \begin{bmatrix} h_1^* r_1^n + h_2 (r_1^{n+1})^* \\ -h_2 (r_1^n)^* + h_1^* r_1^{n+1} \end{bmatrix} = \begin{bmatrix} h_1^* r_1^n \\ h_1^* r_1^{n+1} \end{bmatrix} = (|h_1|^2) \begin{bmatrix} s_n \\ s_{n+1} \end{bmatrix} + \begin{bmatrix} h_1^* w_1^n \\ h_1^* w_1^{n+1} \end{bmatrix} \tag{39}$$

it can still be decoded normally.

Using ML decoding method, the codeword matrix can be written as follows:

$$\hat{s}_n = \arg \min_{\hat{s}_n \in \mathbf{C}} (|h_1|^2 - 1)|\hat{s}_n|^2 + d^2(\tilde{s}_n, \hat{s}_n) \tag{40}$$

$$\hat{s}_{n+1} = \arg \min_{\hat{s}_{n+1} \in \mathbf{C}} (|h_1|^2 - 1)|\hat{s}_{n+1}|^2 + d^2(\tilde{s}_{n+1}, \hat{s}_{n+1}) \tag{41}$$

According to the overlapping coverage area of the SFN systems, two transmitter and one receiver in its overlapping area is reconstruct as MISO systems. Since the two transmitting stations are far enough apart, the channels between the two transmitters and the receiver are independent of each other. Based on this, the parallel Alamouti spatial multiplexing scheme is considered for the reconstruction of SFN system.

According to (29), in a continuous period of two symbols, the transmitting sequence of Tx1 is $[s_1 \quad s_2]$, and the transmitting sequence of Tx2 is $[-s_2^* \quad s_1^*]$.

5. Results

This section uses simulations to validate the system design guidelines obtained in Section 4 and make comparisons with traditional SFN scheme. In this experiment, binary numbers are randomly generated as the transmitting bit stream, and symbol symbols are obtained by QPSK modulation. The parallel Alamouti orthogonal spatial multiplexing in (29) is carried out for one set of modulated symbols, and the traditional SFN dedign scheme in (25) is carried out for another one set of modulated symbols. Assume that the receiver can perfectly estimate the CSI, and ML decoding is used to the SFN receiver.

Figure 5 compares the BER-SNR curves of the parallel Alamouti-SFN scheme and traditional SFN scheme in overlapping coverage area under the same channel conditions.

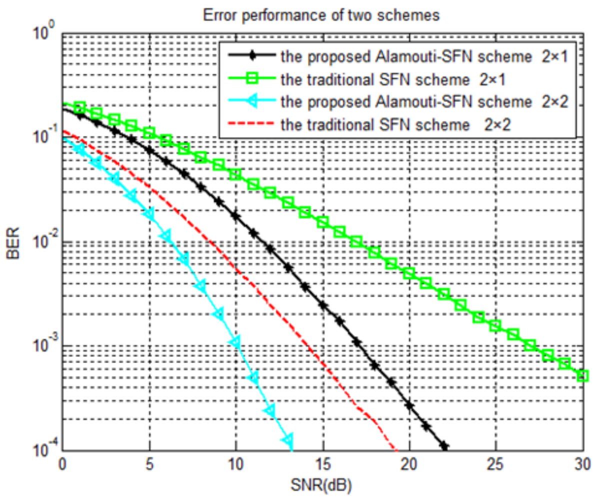

Figure 5. The error performance of parallel Alamouti-SFN scheme and traditional SFN scheme in overlapping coverage area

From figure 5, for one receiving antenna, the introduced parallel Alamouti-SFN scheme can achieve 9.5dB SNR gains when BER is 10^{-3} compared with the traditional SFN scheme; For two receiving antenna, the introduced parallel Alamouti-SFN scheme can achieve 4dB SNR gains when BER is 10^{-3} compared with the traditional SFN scheme.

It is verified that the orthogonal spatial multiplexing scheme can acquire maximum spatial diversity gain when $rN_R = 4$, and the SNR required for the same BER is much lower than other schemes.

Figure 6 compares the BER-SNR curves of the non-overlapping coverage area for the two SFN schemes under the same channel conditions and receiving power.

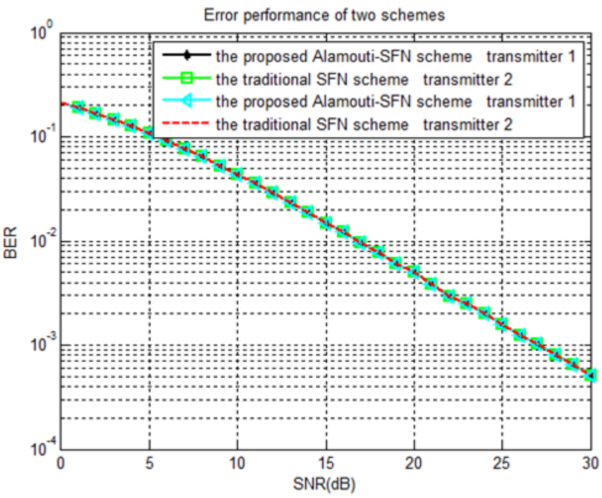

Figure 6. The error performance of parallel Alamouti-SFN scheme and traditional SFN scheme in non-overlapping coverage area

If all conditions being equal, the independent base station in non-overlapping coverage area for the same SFN system can obtained approximate BER performance. Through the vertical comparison with the parallel Alamouti-SFN and the traditional SFN system, the BER curves for the independent base station in non-overlapping coverage area are also approximatively overlapping. Now therefore, the parallel Alamouti-SFN scheme can ensure BER performance in the non-overlapping coverage area, and acquire maximal transmit spatial diversity gain in overlapping coverage area.

6. Conclusions

In this paper, an improved and high-quality spatial multiplexing scheme for SFN system is proposed. Parallel Alamouti orthogonal spatial multiplexing can be reconstruct to promote the spatial freedom, obtain the transmit diversity gain and overcome the deterioration of signal quality in SFN overlapping coverage area.

Simulation results reveal 9.5 dB SNR gains can be achieved when BER is 10-3 compared with the traditional SFN scheme.

It is verified that the proposed spatial multiplexing can effectively utilizes the self-interference of overlapping coverage areas to improve signal quality.

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