Maximum Ratio Diversity Combining Receiver Using Single Radio Frequency Chain and Single Matched Filter

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***Abstract*—We present a new maximum ratio combining (MRC) technique that performs the maximum ratio combining at the radio frequency (RF) level, thereby requiring only one RF chain and one matched filter. We show that it can achieve the same performance as the conventional MRC technique that performs the combining at the baseband (post-detection), which requires multiple RF chains and multiple matched filters. The proposed approach can be extended to multiple input and multiple output (MIMO) system and significantly reduce the hardware complexity. *Index Terms*—Maximum Ratio Combining, Radio Frequency, Matched Filter**

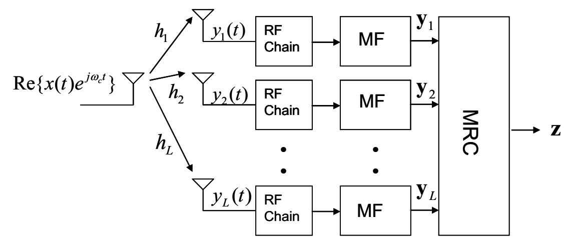

Fig. 1. Conventional post-detection MRC receiver.

I. Introduction

Diversity combining offers a great potential for radio link performance improvement to many of the current and future wireless communications systems. It consists of receiving redundantly the same information-bearing signal over two or more fading channels, and then combining these multiple replicas at the receiver to increase the overall received signal-to-noise ratio (SNR). These multiple replicas can be obtained in space (antenna), frequency, angle, polarization, multipath, and time domains [1]-[3].

This paper is concerned with achieving spatial diversity by using multiple transmit and/or receive antennas. There are several types of combining techniques, among which the maximum ratio combining (MRC) uses each of the multiple replicas in a co-phased and weighted manner such that the highest SNR is achieved at the receiver.

For a diversity system with L diversity branches, the conventional post-detection MRC system consists of L radio frequency (RF) chains (low noise amplifier, frequency converter, and A/D converter) to down convert to baseband, L correlators or matched filter (MF) detectors, and a maximum ratio combiner as illustrated in Fig. 1. A natural concern in implementing such systems is the hardware complexity and size associated with the multiple RF chains and MF detectors. To address this problem, several antenna selection techniques have been proposed [4]-[6]. However, they suffer from performance degradation because they do not use all of the available information.

In this paper, we present a new MRC technique that requires only one RF chain and one MF by performing the MRC task at the RF level. We show that the proposed technique can provide the same performance as the conventional post-detection MRC technique that requires multiple RF chains and multiple MFs. We also show that the proposed technique can be extended to multiple transmit antenna systems. For multiple-input multiple-output (MIMO) systems with K transmit antennas and L receive antennas, the number of RF chains can be reduced from L to K, if $L > K$. For space-time block coded systems, the number of RF chains and matched filters can be reduced from L to 1.

Notation: We use letter s to denote transmitted discrete-time symbols, x to denote transmitted continuous-time signals, y received signals, w signals "weighted" by channel gains, z signals "combined" from multiple diversity branches. Bandpass signals will be decorated with tilde, e.g., $\tilde{y}(t)$. Matched filter outputs will be decorated with bar, e.g., $\bar{z}$. Hilbert transform of a bandpass signal $\tilde{y}(t)$ will be denoted as $\hat{\tilde{y}}(t)$.

II. System Model

A. Single Input Multiple Output

We consider first a single-input multiple-output (SIMO) system composed of one transmit antenna element and L receive antenna elements. Extension to MIMO system will be discussed in Section IV.

We assume that the modulation is linear. The transmitted signal can be written as

$$x(t) = \sum_{i=0}^{\infty} s_i \, p(t - iT) \tag{1}$$

where T is symbol duration, s_i is the i-th transmitted symbol, and $p(t)$ is the spectral shaping pulse. Let "$\star$" denote convolution. We assume that $p(t) \star p(-t)$ satisfies Nyquist criterion

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for inter-symbol interference free transmission:

$$\int_{-\infty}^{\infty} p(t)p(t-iT) = \delta[i] \tag{2}$$

where $\delta[i]$ is the discrete-time delta function. For example, $p(t)$ could be the square-root raised-cosine pulse.

Each channel is assumed to be frequency-non-selective and slowly fading, and the fading processes among the L diversity channels are assumed to be statistically independent. The signal in each channel is corrupted by zero-mean white Gaussian noise. The noise processes in the L channels are assumed to be statistically independent, with identical autocorrelation functions. Thus, the received RF signals for the L channels can be expressed in the form

$$\tilde{y}_l(t) = Re\{y_l(t)e^{j2\pi f_c t}\}, \tag{3}$$

$$y_l(t) = h_l x(t) + n_l(t), \quad l = 1, 2, ..., L \tag{4}$$

where h_l represents the complex channel gain between the transmit antenna element and the l-th receive antenna element, $x(t)$ is the equivalent low-pass transmitted signal, $n_l(t)$ is the low-pass equivalent complex white Gaussian noise at the l-th receive antenna, and f_c is the carrier frequency.

Under the Nyquist assumption (2) and frequency-flat channel assumption, we can do symbol-by-symbol processing at the receiver. We will therefore ignore the symbol index i and consider transmission and reception over a one-symbol duration.

B. Multiple Input Multiple Output

For the case of K transmit antennas and L receive antennas, the channel can be modeled as a matrix whose (l, k) entry $h_{l,k}$ is the gain from the k-th transmit antenna to the l-th receive antenna, $k = 1, \ldots, K$, and $l = 1, \ldots, L$.

The signal received at the l-th receive antenna can be written as

$$\tilde{y}_l(t) = Re\left\{\left[\sum_{k=1}^{K} h_{l,k}x_k(t) + n_l(t)\right] e^{j2\pi f_c t}\right\},\ l = 1, \ldots, L \tag{5}$$

where $x_k(t)$ is the low-pass equivalent signal sent by the k-th transmit antenna, $n_l(t)$ is the low-pass equivalent noise at the l-th receive antenna. We will assume that the channel gains $\{h_{l,k}\}$ are available at the receiver for diversity combining purposes. The channel estimation issue will be discussed in Section V.

III. Proposed MRC for SIMO Systems

We first consider MRC for SIMO systems. MRC for MIMO systems will be discussed in Section IV.

A. Conventional Post-Detection MRC

Consider the channel model in (3). With reference to the conventional MRC system in Fig. 1, the MF output is

$$\bar{y}_l = h_l s + \bar{n}_l, \ \ l = 1, 2, ..., L \tag{6}$$

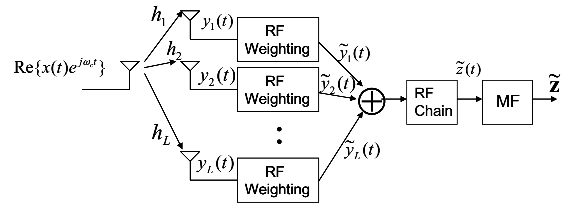

Fig. 2. Proposed MRC receiver.

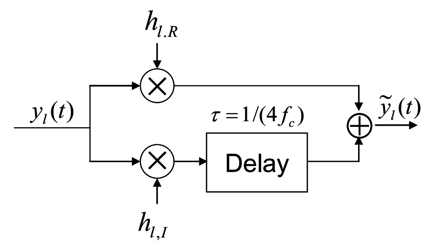

Fig. 3. RF weighting with delay line.

where s and $\bar{n}_l$ are the transmitted data and the discretized noise at MF output, respectively. The MRC output is then

$$z = \sum_{l=1}^{L} h_l^* \bar{y}_l = \sum_{l=1}^{L} [|h_l|^2 s + h_l^* \bar{n}_l]. \tag{7}$$

Such conventional MRC system requires L RF chains and L matched filters. In the following, we propose a combining scheme then requires only one RF chain and one matched filter.

B. Proposed MRC

The block diagram of the proposed MRC receiver is shown in Fig. 2. The RF weighting block for the l-th receive antenna (shown in Fig. 3) produces

$$\begin{aligned}\tilde{w}_l(t) &= h_{l,R}\tilde{y}_l(t) + h_{l,I}\tilde{y}_l(t-\tau)\\ &= Re\{[h_{l,R}y_l(t) + h_{l,I}y_l(t-\tau)e^{-j2\pi f_c \tau}] \cdot e^{j2\pi f_c t}\}\end{aligned}$$

where $h_{l,R} = Re\{h_l\}$ and $h_{l,I} = Im\{h_l\}$. If the time delay τ is chosen to be

$$\tau = 1/(4f_c) \tag{8}$$

then we obtain $x(t-\tau) \simeq x(t)$ and $e^{-j2\pi f_c \tau} = -j$. It should be noted that $x(t)$ and $x(t-\tau)$ are virtually the same in practice. To see how good the approximation is, consider the case f_c is 1GHz and the symbol duration T is 10^{-6} second, corresponding to the symbol rate of 1M symbols per second. Then, τ/T is 2.5×10^{-4}, which is negligibly small. Similarly, we also have $n_l(t) \simeq n_l(t-\tau)$, cf. (4). Hence,

$$\begin{aligned}\tilde{w}_l(t) \ &= \ Re\{[h_{l,R}h_l x(t) - jh_{l,I}h_l x(t-\tau)\\ &\quad +h_{l,R}n_l(t) - jh_{l,I}n_l(t-\tau)]e^{j2\pi f_c t}\}\\ &\simeq \ Re\{[|h_l|^2 x(t) + h_l^* n_l(t)]e^{j2\pi f_c t}\}.\end{aligned} \tag{9}$$

Summing the weighted signals $\{\tilde{w}_l(t)\}$ and down converting the summation to the baseband yield a combined baseband signal

$$z(t) \simeq \sum_{l=1}^{L} \left[|h_l|^2 x(t) + h_l^* n_l(t) \right]. \tag{10}$$

Then, the MF output is

$$\bar{z} \simeq \sum_{l=1}^{L} [|h_l|^2 s + h_l^* \bar{n}_l]. \tag{11}$$

Comparing (7) and (11), we see that the proposed MRC receiver (Fig. 2) that uses single RF chain and single MF produces essentially the same output as does the conventional MRC receiver (Fig. 1) that requires L RF chains and L MFs. The reduction in hardware complexity becomes more significant for larger L.

C. Hilbert Transform Viewpoint

In this subsection we show that the delay line can be interpreted as a Hilbert transformer. Consider a generic bandpass (RF) signal

$$\tilde{v}(t) = Re\{v(t)e^{j2\pi f_c t}\} = Re\{[v_R(t) + jv_I(t)]e^{j2\pi f_c t}\}, \tag{12}$$

where $v(t) = v_R(t) + jv_I(t)$ is the low-pass equivalent of $\tilde{v}(t)$. The Hilbert transform of $v_R(t)\cos(2\pi f_c t)$ is

$$\mathcal{F}^{-1}\left\{\frac{(-j)V_R(f-f_c) + jV_R(f+f_c)}{2}\right\} = v_R(t)\sin(2\pi f_c t)$$

where $\mathcal{F}^{-1}$ denotes inverse Fourier transform, and $V_R(f)$ is the Fourier transform of $v_R(t)$. Similarly, we can show that the Hilbert transform of $v_I(t)\sin(2\pi f_c t)$ is $-v_I(t)\cos(2\pi f_c t)$. Since the Hilbert transform is linear, it follows that the Hilbert transform $\hat{\tilde{v}}(t)$ of $\tilde{v}(t)$ is

$$\begin{aligned} \mathcal{H}\{\tilde{v}(t)\} &= v_R(t)\sin(2\pi f_c t) + v_I(t)\cos(2\pi f_c t) && (13)\\ &= Re\{[v_R(t) + jv_I(t)]\cdot(-j)e^{j2\pi f_c t}\} && (14)\\ &\simeq \tilde{v}(t - \frac{1}{4f_c}) && (15) \end{aligned}$$

where the approximation in (14) is good if $v(t)$ is a low-pass signal and f_c is much larger than the bandwidth of $v(t)$ such that $v(t - 1/(4f_c)) \simeq v(t)$. Equations (13)-(15) show that the Hilbert transform of a bandpass signal is equivalent to multiplying the low-pass equivalent signal by $-j$, and can be implemented approximately by delaying the signal in time by $1/(4f_c)$.

With this viewpoint, we can write the output of the RF weighting system (Fig. 3) as

$$\tilde{w}_l(t) \simeq h_{l,R}\tilde{y}_l(t) + h_{l,I}\hat{\tilde{y}}_l(t) = Re\{h_l^* y_l(t)e^{j2\pi f_c t}\}, \tag{16}$$

which is the desired combining in the RF domain.

IV. RF COMBINING IN MIMO SYSTEMS

A. MRC in General MIMO System

Consider now the MIMO channel model in (5). Assume that the noise vector is spatially uncorrelated. We can apply spatial matched filtering to the signals from the L receive antennas, to obtain

$$z_k(t) = \sum_{l=1}^{L} h_{l,k}^* \cdot y_l(t), \quad k = 1, \ldots, K \tag{17}$$

in the low-pass domain [cf. (5)]. The corresponding RF signals can be written as follows [cf. (13)–(15)]:

$$\begin{aligned} \tilde{z}_k(t) &= Re\{z_k(t)e^{j2\pi f_c t}\}, \quad \text{for } k = 1, \ldots, K \\ &= \sum_{l=1}^{L}\left\{Re[h_{l,k}]\tilde{y}_l(t) + Im[h_{l,k}]\hat{\tilde{y}}_l(t)\right\}. && (18) \end{aligned}$$

As in the SIMO case in Section III, the Hilbert transforms can be implemented in delay lines (Fig. 3), hybrid power dividers, or RLC components.

Following the standard argument that leads to the sufficiency of sampled matched filter output, e.g., [7], it can be shown that the RF combining in (18) produces sufficient statistics [8], and does not lead to a loss of information as compared to $\tilde{y}_l(t),\ l = 1, \cdots, L$. Therefore, the benefit of the RF combining is as follows: when $L > K$, the RF combining in (18) results in less number of RF signals to be converted to the baseband. For example, if $K = 2$ and $L = 4$, we need 4 RF chains to down convert $\{\tilde{y}_l(t), l = 1, 2, 3, 4\}$ without the RF combining. With the RF combining in (18), we only need 2 RF chains to down convert $\tilde{z}_k(t)$, $k = 1, 2$.

When $K \geq L$, the RF combining does not offer a saving in the number of RF chains. We also comment that for general MIMO transmissions, it is in general impossible to use only one RF chain to preserve the useful information that is received from the multiple receive antennas. The reason is that the using one RF chain, we can only obtain one complex symbol per symbol duration, whereas multiple dimensions is generally needed to process the received signals. One exception, however, is when the transmitter uses a beamforming mode, which we discuss in the following.

B. MRC in Beamforming Systems

In a MIMO system, if the multiple transmit antennas are used in the beamforming mode, which means that the transmitted signals at all transmit antennas are multiples of each others, then RF combining at the receiver is possible and the combined signal only needs one RF chain to be down converted.

Let q_k denote the complex weights used at the transmitter. The vector $[q_1, \ldots, q_K]^T$ is often called the beamforming vector. The transmitted signal at the k-th antenna is given by $Re[q_k x(t)e^{2\pi f_c t}]$, where $x(t)$ is the low-pass equivalent information-bearing signal. The received signal at the l-th antenna is

$$\tilde{y}_l(t) = Re\left\{y_l(t)e^{j2\pi f_c t}\right\}, \quad l = 1, \ldots, L \tag{19}$$

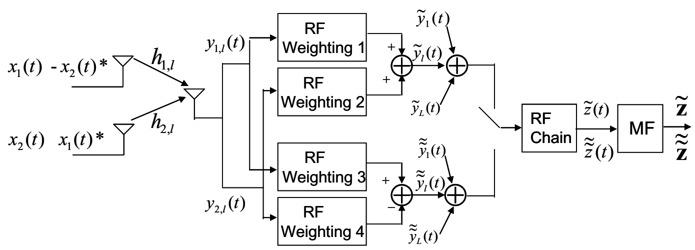

Fig. 4. Proposed MRC receiver for space-time coded signal.

where

$$y_l(t) = \sum_{k=1}^{K} [h_{l,k} q_k x(t) + n_l(t)]. \tag{20}$$

We can combine the signals in the following way

$$\tilde{z}(t) = \sum_{l=1}^{L} \left\{ Re\left[\sum_{k=1}^{K} h_{l,k} q_k\right] \tilde{y}_l(t) + Im\left[\sum_{k=1}^{K} h_{l,k} q_k\right] \hat{\tilde{y}}_l(t) \right\}. \tag{21}$$

This RF combining implements a spatial matched filter operation and does not lead to a loss of information. Since there is only RF signal after combining, only one RF chain is needed to down convert $\tilde{z}(t)$ to the baseband for further processing without loss of information.

It can be shown that

$$\begin{aligned} \tilde{z}(t) &= Re\left\{ \sum_{l=1}^{L} \left[\left| \sum_{k=1}^{K} h_{l,k} q_k \right|^2 x(t) \right.\right. \\ &\left.\left. + \left(\sum_{k=1}^{K} h_{l,k} q_k \right)^* n_l(t) \right] e^{j2\pi f_c t} \right\}. \end{aligned} \tag{22}$$

Let $N_0 = E[|n_l(t)|^2]$, then the SNR of $\tilde{z}(t)$ can be shown to be

$$\mathrm{SNR} = \sum_{l=1}^{L} \left| \sum_{k=1}^{K} h_{l,k} q_k \right|^2 \cdot E\left[\, |x(t)|^2 \,\right] / N_0. \tag{23}$$

A diversity order of KL can be provided if the beamforming vector is chosen judiciously according to the singular value decomposition of the MIMO channel.

C. MRC in Space-Time Block Coded Systems

In this subsection we consider a MIMO system with space-time block coded transmission. We will consider a transmitter with two antennas elements, sending Alamouti-type space-time block coded signals. Extension to more than two transmit antenna elements is straightforward. The receiver has L receive antenna elements for maximum ratio combining.

The system block diagram for the proposed RF combining scheme is shown in Fig. 4. At a given signaling interval, signals $x_1(t)$ and $x_2(t)$ are transmitted simultaneously from antenna 1 and 2, respectively. In the following we show that the proposed MRC scheme shown in Fig. 4, requiring single RF chain and single MF, can provide diversity order of $2L$.

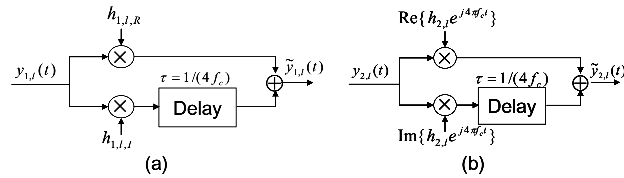

Fig. 5. (a) RF weighting 1, (b) RF weighting 2.

The received RF signals at the l-th receive antenna, $\tilde{y}_{l,1}(t)$ and $\tilde{y}_{l,2}(t)$, in two consecutive symbol intervals are given by

$$\begin{aligned} \tilde{y}_{l,1}(t) &= Re\{[h_{l,1}x_1(t) + h_{l,2}x_2(t) + n_{l,1}(t)]e^{j2\pi f_c t}\} \\ \tilde{y}_{l,2}(t) &= Re\{[-h_{l,1}x_2^*(t) + h_{l,2}x_1^*(t) + n_{l,2}(t)]e^{j2\pi f_c t}\} \end{aligned}$$

where $h_{l,i}$, $i = 1, 2$, is the complex Gaussian channel gain between the i-th transmit antenna and the l-th receive antenna, and $n_{l,2}(t)$ and $n_{l,2}(t)$ are the complex Gaussian noise at the l-th receive antenna.

The output of RF weighting 1 block, shown in Fig. 5(a), is

$$\tilde{w}_{l,1}(t) \simeq Re\{[h_{l,1}^* h_{l,1} x_1(t) + h_{l,1}^* h_{l,2} x_2(t) + h_{l,1}^* n_{l,1}(t)] e^{j2\pi f_c t}\} \tag{24}$$

Similarly, the output of RF weighting 2 block, shown in Fig. 5(b), is

$$\begin{aligned} \tilde{w}_{l,2}(t) &\simeq Re\{[(h_{l,2}e^{j4\pi f_c t})^*(-h_{l,1}x_2^*(t) + h_{l,2}x_1^*(t)) \\ &\quad + n_{l,2}(t)] \cdot e^{j2\pi f_c t}\} \\ &= Re\{[-h_{l,1}^* h_{l,2} x_2(t) + h_{l,2}^* h_{l,2} x_1(t) \\ &\quad + h_{l,2} \cdot n_{l,2}^*(t)] \cdot e^{j2\pi f_c t}\} \end{aligned} \tag{25}$$

Adding $\tilde{w}_{l,1}(t)$ and $\tilde{w}_{l,2}(t)$ yields

$$\begin{aligned} \tilde{w}_l(t) &= \tilde{w}_{l,1}(t) + \tilde{w}_{l,2}(t) \\ &\simeq Re\{[(|h_{l,1}|^2 + |h_{l,2}|^2)x_1(t) + h_{l,1}^* n_{l,1}(t) \\ &\quad + h_{l,2} n_{l,2}^*(t)] \cdot e^{j2\pi f_c t}\}. \end{aligned} \tag{26}$$

Summing the weighted signals $\{\tilde{w}_l(t)\}$ and down converting the summation to the baseband yields

$$z(t) \simeq \sum_{l=1}^{L} [(|h_{l,1}|^2 + |h_{l,2}|^2)x_1(t) + h_{l,1}^* n_{l,1}(t) + h_{l,2} n_{l,2}^*(t)]. \tag{27}$$

We remind the reader that the approximation in (27) is excellent for almost all practical communication systems (except for ultra-wideband systems).

The noise terms in (27) are complex Gaussian with mean zero and variance $(|h_{l,1}|^2 + |h_{l,2}|^2)N_0$, providing the SNR

$$\mathrm{SNR}_1 = \sum_{l=1}^{L} (|h_{l,1}|^2 + |h_{l,2}|^2) E[\, |x_1(t)|^2 \,] / N_0 \tag{28}$$

which is identical to that of the conventional post-detection MRC technique that uses L RF chains and L MFs.

Similarly, applying $y_{l,1}(t)$ to RF weighting 3 block, shown in Fig. 6(a), yields

$$\begin{aligned} \tilde{w}_{l,3}(t) &\simeq Re\{[|h_{l,2}|^2 x_2(t) + h_{l,1} h_{l,2}^* x_1(t) \\ &\quad + h_{l,2}^* n_{l,1}^*(t)] \cdot e^{j2\pi f_c t}\}, \end{aligned} \tag{29}$$

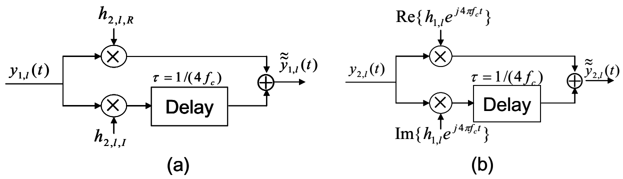

Fig. 6. (a) RF weighting 3, (b) RF weighting 4.

and applying $y_{l,2}(t)$ to RF weighting 4 block, shown in Fig. 6(b), yields

$$\begin{aligned} \tilde{w}_{l,4}(t) &\simeq Re\{[-|h_{l,1}|^2 x_2(t) + h_{l,1}h_{l,2}^* x_1(t) \\ &\quad + h_{l,1}n_{l,2}^*(t)] \cdot e^{j2\pi f_c t}\} \end{aligned} \tag{30}$$

Subtracting $\tilde{w}_{l,4}(t)$ from $\tilde{w}_{l,3}(t)$ yields

$$\begin{aligned} \tilde{w}_l'(t) &= \tilde{w}_{l,3}(t) - \tilde{w}_{l,4}(t) \quad (31) \\ &\simeq Re\{[(|h_{l,1}|^2 + |h_{l,2}|^2)x_2(t)h_{l,2}^* n_{l,1}(t) \\ &\quad - h_{l,1}n_{l,2}^*(t)] \cdot e^{j2\pi f_c t}\}. \quad (32) \end{aligned}$$

Summing the weighted signals $\{\tilde{w}_l'(t)\}$ and down converting the summation to the baseband yields

$$z'(t) \simeq \sum_{l=1}^{L} \left[(|h_{l,1}|^2 + |h_{l,2}|^2)x_2(t) + h_{l,2}^* n_{l,1}(t) - h_{l,1}n_{l,2}^*(t)\right] \tag{33}$$

Therefore, it follows from (33) that the SNR for $x_2(t)$ is given by

$$\mathrm{SNR}_2 = \sum_{l=1}^{L}(|h_{l,1}|^2 + |h_{l,2}|^2)E[\,|x_2(t)|^2\,]/N_0, \tag{34}$$

which shows that the diversity order for $x_2(t)$ is also $2L$.

Reducing the Number of Mixers: In directly implementing the block diagrams of Fig. 5(b) and Fig. 6(b), 4 mixers will be required for each diversity branch, hence requiring $4L$ mixers in total for L diversity branches. However, the number of mixers in RF weighting 2 and 4 blocks can be reduced as follows. The output of RF weighting 2 block can be expressed as

$$\begin{aligned} \tilde{w}_{l,2}(t) &= \tilde{y}_{l,2}(t) \cdot Re\{h_{l,2}e^{j4\pi f_c t}\} - \tilde{y}_{l,2}(t-\tau) \\ &\quad \cdot Im\{h_{l,2}e^{j4\pi f_c (t-\tau)}\} \\ &= \underbrace{[\tilde{y}_{l,2}(t)h_{l,2,R} + \tilde{y}_{l,2}(t-\tau)h_{l,2,I}]}_{=\alpha_l}\cos(4\pi f_c t) \\ &\quad - \underbrace{[\tilde{y}_{l,2}(t)h_{l,2,I} - \tilde{y}_{l,2}(t-\tau)h_{l,2,R}]}_{=\beta_l}\sin(4\pi f_c t) \end{aligned} \tag{35}$$

where $h_{l,2,R} = Re\{h_{l,2}\}$ and $h_{l,2,I} = Im\{h_{l,2}\}$. Then, the summation

$$\sum_{l=1}^{L}\tilde{w}_{l,2}(t) = \left[\sum_{l=1}^{L}\alpha_l\right]\cos(4\pi f_c t) - \left[\sum_{l=1}^{L}\beta_l\right]\sin(4\pi f_c t) \tag{36}$$

only requires two mixers, although it requires more weightings. Similarly, the summation $\sum_{l=1}^{L}\tilde{w}_{l,4}(t)$ requires two mixers. Therefore, the proposed MRC receiver in Fig. 4 requires a total of 4 mixers in RF combining of all L branches. Comparing the overall hardware complexity, the conventional MRC requires L RF chains and L MFs, whereas the proposed MRC requires 1 RF chain, 1 MF, and 4 mixers in space-time block coded systems.

V. Channel Estimation Issue

For SIMO systems, MRC requires knowledge of the channel gains, $h_1, h_2, ..., h_L$. This seems to necessitate the use of L RF chains and L MF detectors. However, in a sufficiently slowly-varying environment, the antenna elements can be multiplexed to the RF chain during the training period. That is, the RF chain is connected to the first antenna element during the first part of the training sequence, then to the second antenna element during the next part, and so on. Thus, we need a few more training symbols and extra training time, not more RF chains. Especially in high data rate systems, those additional training bits decrease the spectral efficiency in a negligible way.

The same idea of switching the receiver RF chain to different receive antennas can be applied to MIMO systems. In the MIMO case, while switched to one receive antenna the channel gains from the multiple transmit antennas to a specific receive antenna are estimated. Again, the price paid in this case is extra training symbols and training time, which can be quite acceptable in high-rate transmissions.

VI. Conclusions

We have proposed a new maximum ratio combining technique that performs the maximum ratio combining at the RF-level and thus requires only one RF chain and one MF. We showed that it can achieve the same performance as the conventional post-detection MRC technique that requires multiple RF chains and multiple MFs. The proposed approach can significantly reduce the hardware complexity and can also be applied in multiple input and multiple output systems.

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