Parallel Multistage Decision Feedback Equalizer for Single-Carrier Layered Space-Time Systems in Frequency-Selective Channels

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Space-time transmission techniques can greatly increase the spectral efficiency. In this paper, a parallel multistage decision feedback equalizer (PMDFE) is proposed for single-carrier layered space-time systems with a fixed cyclic prefix over frequency-selective channels. It is composed of a parallel interference canceller, a multiple-input single-output decision feedback equalizer (MISO-DFE), and a linear combiner. The soft output of the MISO-DFE is linearly combined with the previous tentative soft decision. In addition, an algorithm is proposed to obtain tentative soft and hard decisions for initializing the equalizer. The initializing complexity of the PMDFE is lower than that of MIMO-OFDM. Simulation results show that the PMDFE outperforms MIMO-OFDM and previously existing equalizers for single-carrier layered space-time systems.

Keywords and phrases: MIMO, OFDM, PIC, DFE, single carrier.

1. INTRODUCTION

Conventional technique to mitigate intersymbol interference (ISI) is equalization in time domain (TD). For broadband transmission, the complexity of equalization in TD will be very high. Orthogonal frequency division multiplexing (OFDM) transmission [1] that can be considered as a one-tap linear equalizer is a promising technique to eliminate ISI. However, the high peak-to-average power ratio will degrade its performance. Another technique to eliminate ISI is single-carrier modulation with a cyclic prefix (CP) and frequency domain (FD) equalization. Frequency domain linear equalizers (FDLE) proposed in [2, 3, 4] make this technique a candidate for broadband transmission. Recently, a new decision feedback equalizer (DFE) is proposed in [5, 6] for single-carrier system with a fixed CP in single-input single-output (SISO) frequency-selective channel. Its forward filter is performed in FD and the feedback filter is in TD. A fixed CP is utilized to initialize the feedback filter. This scheme can achieve better performance than the FDLE. Moreover, the complexity of this scheme does not increase with RMS delay spread. Results in [7] show that the fixed CP can be used for channel estimation and synchronization.

For multimedia services, more bandwidth is demanded. Space-time transmission techniques [8, 9] can greatly increase the spectral efficiency. Layered space-time receivers with interference cancellation are proposed [10] to mitigate ISI and cochannel interference (CCI) for frequency-selective channels. However, the complexity of space-time processing in TD increases with RMS delay spread. Multiple-input
multiple-output OFDM (MIMO-OFDM) scheme [11] is a satisfactory solution not only to mitigate ISI but also to obtain high spectral efficiency. The MIMO decision feedback equalizer (MIMO-DFE) [12] with its forward filter performed in FD and feedback filter implemented in TD is proposed for single-carrier layered space-time systems. However, the complexity of the MIMO-DFE is too high to be implemented. Meanwhile, a linear equalizer along with one stage parallel interference cancellation (PIC) in FD (FDLE-PIC) is also proposed in [12]. Since PIC in [12] is equivalent to a DFE in space domain (SD), the FDLE-PIC can be considered as a DFE in SD combined with a linear equalizer in FD.

In this paper, a multiple-input single-output decision feedback equalizer (MISO-DFE) with its forward and feedback filter performed in FD and TD, respectively, is proposed for equivalent single-input multiple-output channels. Then a parallel multistage decision feedback equalizer (PMDFE) is proposed for single-carrier layered space-time systems with a fixed CP. It is composed of a parallel interference canceller, a MISO-DFE, and a linear combiner. The soft output of MISO-DFE is linearly combined with the previous tentative soft decision. The proposed PMDFE is equivalent to the DFE in SD combined with a MISO-DFE in FD and TD. Moreover, partial parallel interference cancellation (PPIC) idea stemming from multiuser detector [13] is introduced to reduce the effect of error propagation. The complexity of the PMDFE does not increase with RMS delay spread and the performance of the PMDFE is better than the MIMO-OFDM and the FDLE-PIC. In addition, an algorithm is proposed to obtain tentative soft and hard decisions for the initialization of the PMDFE.

The paper is organized as follows. In Section 2, channel models and single-carrier layered space-time systems are introduced. Then the PMDFE is derived in Section 3. In Section 4, we formulate an algorithm to obtain tentative soft and hard decisions for the initialization of the PMDFE. In Section 5, the complexity of existing schemes is compared in terms of complex multiplication. Finally, in Section 6, the performance of the proposed equalizer is compared with existing schemes through computer simulations.

2. CHANNEL MODEL AND SINGLE-CARRIER SYSTEMS

The layered space-time architecture is given in Figure 1. As shown in Figure 2, the $k$th transmission block from the $i$th Tx is

$$Q_i[k] = [s_i[k(T + L)], s_i[k(T + L) + 1], \ldots, s_i[(k + 1)(T + L) - 1]]$$

$$= [d_i[kT], d_i[kT + 1], \ldots, d_i[kT + T - 1], \text{PN}_{i,0}, \text{PN}_{i,1}, \ldots, \text{PN}_{i,L-1}], \quad \text{for } i = 1, 2, \ldots, M,$$

(1)

where $L$ is the length of CP, the last $L$ symbols are pseudonoise (PN) sequence, and $T$ denotes the number of transmitted symbols in one data block. These data blocks are simultaneously transmitted by $M$ transmitters. In addition, the same PN sequence is transmitted before the first data block. This technique that forces linear convolution into circular convolution was first proposed in [14]. PN sequences are fixed, known at the receiver, and suitable for use with a DFE. The conventional technique [1] is to add a CP in front of a data block. However, the CP of the conventional technique changes with data block. The received signal from the $q$th receive antenna at time instant $n$ is

$$r_q[k(T + L) + n] = \sum_{i=1}^{M} \sum_{l=0}^{L_{iq}-1} s_i[k(T + L) + n - l] h_{iq}[l] + v_q[k(T + L) + n],$$

$$n = -L, -(L-1), \ldots, T + L - 1, \quad q = 1, \ldots, N,$$

(2)

where $L_{iq}$ is the memory length of the channel from the $i$th transmit antenna to the $q$th receive antenna, $h_{iq}[l]$ is the $l$th tap of the channel corresponding to the $i$th transmit antenna and the $q$th receive antenna, and $v_q[n]$ denotes the complex additive white Gaussian noise with variance $\sigma_q^2$ at the $q$th receive antenna. For convenience, we will drop the index $k$ in the following discussion. Discarding the CP, the $(T + L)$-sized DFT of the received signal at the $q$th receive antenna is

$$R_q[f] = \sum_{i=1}^{M} S_i[f] H_{iq}[f] + V_q[f], \quad f = 0, 1, \ldots, T + L - 1,$$

(3)

where

$$S_i[f] = \sum_{n=0}^{T+L-1} s_i[n] e^{-j2\pi nf/(T+L)},$$

(4)

$$H_{iq}[f] = \sum_{l=0}^{L_{iq}} h_{iq}[l] e^{-j2\pi lf/(T+L)}.$$
In the following section, an algorithm will be introduced to make decisions for the transmitted symbols have been obtained. The interference to the transmitted signal for the kth substream in FD at the fth subchannel can be expressed as follows:

$$\mathbf{R}[f] = \mathbf{R}[f] - \sum_{t=1}^{M} \hat{s}_t^{-1}[f](\mathbf{H}[f]),$$

where (\(\mathbf{H}[f]\)) is the rth column of channel matrix and \(\hat{s}_t^{-1}[f]\) is the DFT of the tentative hard decision \(\hat{s}_t^{-1}[n]\) for the rth substream at the \((j-1)\)th stage. As shown in Figure 3, applying the weight vector of the forward filter to the modified received signal \(\mathbf{R}[f]\),

$$Y_i[f] = \mathbf{W}_f^H[f]\mathbf{R}[f],$$

where \((\cdot)^H\) denotes the conjugate transpose operation. With the coefficient \(b_{l,i}^{\text{FBF}} = [b_{l,i}^{\text{FBF}}, \ldots, b_{l,i}^{\text{FBF}}]^T\) of the feedback filter, the soft output of the MISO-DFE for the i th substream at the f th stage is

$$m_i^j[n] = y_i[n] + \sum_{l=1}^{L} b_{l,i}^{\text{FBF}}\hat{s}_l^{-1}[n-l], \quad n = 0, 1, \ldots, T - 1,$$

where \(y_i[n]\) is the inverse DFT of \(Y_i[f]\) and the feedback signal \(\hat{s}_l^{-1}[n]\) is defined as

$$\hat{s}_l^{-1}[n] = \begin{cases} P_{N_l,-n}, & -L \leq n \leq -1, \\ d_{l}^{-1}[n], & 0 \leq n \leq T - 1. \end{cases}$$
The mean square error of theith substream is
\[ J_{i,\text{MMSE}} = E\left\{ \left| m^i[n] - d[n] \right|^2 \right\} \].

Assuming the previous tentative hard decisions are all correct, the mean square error can be expressed as follows:
\[ J_{i,\text{MMSE}} = E\left\{ \left| y_i[n] + \sum_{l=1}^{L} b_{i,l}^{\text{BBF}} s_i[n-l] - s_i[n] \right|^2 \right\} \].

It can be expressed in FD as
\[ J_{i,\text{MMSE}} = \frac{1}{T+L} \sum_{f=0}^{T+L-1} \left\{ \sigma_d^2 |W_i^H[f](H[f])_i + B_i[f] - 1|^2 + \sigma_d^2 |W_i^H[f]W_i[f]| \right\} \]
(13)

where \(B_{i,\text{BBF}}[f]\) is the DFT of \(b_{i,\text{BBF}}\) and \(\sigma_d^2\) denotes the power of the transmitted symbol. Applying the gradient method to (13), we obtain
\[ \frac{\partial J_{i,\text{MMSE}}}{\partial W_i[f]} = 0 \].

Therefore, the weight vector of the forward filter can be expressed as follows:
\[ W_i[f] = \left[ (H[f])_i (H[f])_i^H + \frac{\sigma_d^2}{\sigma_v^2} I \right]^{-1} (H[f])_i (1 - B_{i,\text{BBF}}[f]) \],
(15)

where \(^*\) denotes the conjugate operation and \(I\) denotes an identity matrix. With the matrix inversion lemma in (A.2), the weight vector of the forward filter can be computed with low complexity when \(B_{i,\text{BBF}}[f]\) is known. From (13) and (15), the mean square error can be rewritten as follows:
\[ J_{i,\text{MMSE}} = \frac{1}{T+L} \sigma_d^2 \sum_{f=0}^{T+L-1} \frac{|1 - B_{i,\text{BBF}}[f]|^2}{\sigma_v^2/\sigma_d^2 + ||(H[f])_i||^2} \],
(16)

where \(\|\cdot\|\) denotes the Frobenius norm. The derivation of (16) is given in the appendix. Applying the gradient method to (16), the optimum coefficient \(b_{i,\text{BBF}}\) of the feedback filter in the MMSE sense is
\[ b_{i,\text{BBF}} = R_i^{-1} a_i \].
Then, the estimate \( Y_0^0[f] = [Y_0^0[f], \ldots, Y_M^0[f]]^T \) of each substream in FD at the \( f \)th subchannel is

\[
Y_0^0[f] = (W_0^0[f])^H R[f].
\] (22)

Therefore, the estimate for the \( i \)th substream at the \( f \)th subchannel is

\[
Y_i^0[f] = G_i^0[f] S_i[f] + V_i^0[f],
\] (23)

where

\[
G_i^0[f] = (W_0^0[f])^H_i (H[f]),
\]

\[
V_i^0[f] = \sum_{t=1, t\neq i}^M [(W_0^0[f])_{t}^H (H[f])_{t} + (W_0^0[f])_{t}^H V[f].
\] (24)

The colored noise \( V_i^0[f] \) will degrade the performance of the DFE. Therefore, a prewhitening filter is required for each substream in FD. The output of the prewhitening filter is

\[
\hat{Y}_i^0[f] = \frac{Y_i^0[f]}{\sum_{t=1, t\neq i}^M \sigma_d^2_t \left( (W_0^0[f])_{t}^H (H[f])_{t} \right)^2 / \sigma_d^2 + \left\| (W_0^0[f])_{t} \right\|^2}
\]

\[
= G_i^0[f] S_i[f] + \hat{V}_i^0[f],
\] (25)

where the equivalent channel frequency response \( \tilde{G}_i^0[f] \) is defined as

\[
\tilde{G}_i^0[f] = \frac{G_i^0[f]}{\sum_{t=1, t\neq i}^M \sigma_d^2_t \left( (W_0^0[f])_{t}^H (H[f])_{t} \right)^2 / \sigma_d^2 + \left\| (W_0^0[f])_{t} \right\|^2},
\] (26)

and the interference plus noise \( \hat{V}_i^0[f] \) is

\[
\hat{V}_i^0[f] = \frac{V_i^0[f]}{\sum_{t=1, t\neq i}^M \sigma_d^2_t \left( (W_0^0[f])_{t}^H (H[f])_{t} \right)^2 / \sigma_d^2 + \left\| (W_0^0[f])_{t} \right\|^2}.
\] (27)

With the same derivation as in [5], the optimum coefficients of the DFE for the equivalent SISO frequency-selective channel can be computed with low complexity, and initial tentative hard and soft decisions can be obtained.

### 5. COMPLEXITY EVALUATION

In this section, the complexity of existing schemes is compared in terms of complex multiplications. The data block size of MIMO-OFDM is \( T \) and the length of the prefix of MIMO-OFDM is \( B \) is the span of the feedback filter of the MIMO-DFE. \( L_s \) denotes the number of the total detection stages of the PMDFE.

From Table 1, we can see that the initialization complexity of the MIMO-DFE [10] is too high to be implemented.

| Table 1: Complexity comparison of the equalizers design for MIMO frequency-selective channel in the case \( M = N \). |
|-------------------|-------------------|-----------------------------|
| MIMO-OFDM         | \((2M^4 + 2M^3)T\) | Null                        |
| SC-MIMO-DFE       | \((T + L)^3M^4/3 + 2M^4(T + L)^2 + M^4L\) | \(ML(T + L)/2(\log_2(T + L) - 1)\) |
| SC-FDLE           | \(7(T + L)M^4/3\)  | Null                        |
| SC-FDLE-PIC       | \(7M^4/3(T + L) + M^2(T + L)\) | Null                        |
| SC-PMDFE          | \(17M^4(T + L) + 6ML^2\) | \(M(T + L)(\log_2(T + L) - 1)\) |

| Table 2: Complexity of signal detection for MIMO frequency-selective channel in the case \( M = N \). |
|-------------------|-------------------|-----------------------------|
| MIMO-OFDM         | \(MT/2(\log_2(T) - 1)\) | Null                        |
| SC-MIMO-DFE       | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) |
| SC-FDLE           | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) |
| SC-FDLE-PIC       | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) |
| SC-PMDFE          | \((L + 1)M(T + L)\) \(\times (\log_2(T + L) - 1)\) | \((L + 1)M(T + L)\) \(\times (\log_2(T + L) - 1)\) |

FFT | IFFT | PIC/PPIC | FFF | FBF
---|-----|---------|-----|-----
MIMO-OFDM | \(MT/2(\log_2(T) - 1)\) | Null | Null | \(MT\) | Null
SC-MIMO-DFE | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | Null | \(M^2(T + L)\) | \(LM^2(T + L)\)
SC-FDLE | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | Null | \(M^2(T + L)\) | Null
SC-FDLE-PIC | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | \(M(T + L)/2\) \(\times (\log_2(T + L) - 1)\) | Null | \(M^2(T + L)\) | Null
SC-PMDFE | \((L + 1)M(T + L)\) \(\times (\log_2(T + L) - 1)\) | \((L + 1)M(T + L)\) \(\times (\log_2(T + L) - 1)\) | \(L_sM^2(T + L)^3\) \(\times (\log_2(T + L) - 1)\) | \((L + 1)^2M^2(T + L) + +M(T + L)\) | \(TLML + LMT\)
The initialization complexity of the MIMO-OFDM is also very high and is mainly due to the expected SINR ordering.

From Table 2, it can be observed that the complexity of signal detection in various schemes mainly comes from the FFT and IFFT operations. Simulation results in the next section show that satisfactory performance can be achieved through three stages. MIMO-OFDM system has the lowest complexity of signal detection among these schemes.

### 6. SIMULATION RESULTS

The performance of various schemes is evaluated for QPSK modulation over a ten-tap channel in Table 3. Each tap is independent Rayleigh faded and quasistatic. There is no channel coding. The bit error rates (BERs) of various schemes are given versus the averaged received SNR per receiver antenna. The perfect channel information is assumed to be known at the receiver.

In Figure 5, the performance of the PMDFE with partial decision feedback parameters $p_1 = 0.7$, $p_2 = 0.8$, $p_3 = 0.9$, and $p_4 = 0.95$ is given for space-time architecture (2Tx, 2Rx). If the target BER is $10^{-2}$, at the first stage about 2.0 dB gain can be obtained over the initial stage. Through, four stages, 2.7 dB gain can be obtained compared to the initial stage. From Figure 5, we can see that satisfactory performance can be achieved through three stages.

In Figure 6, the performance comparison of the PMDFE with different partial decision feedback parameters is presented for (2Tx, 2Rx). When partial decision feedback parameters are $p_1 = 0.7$, $p_2 = 0.8$, $p_3 = 0.9$, and $p_4 = 0.95$, the PMDFE is robust against error propagation. Therefore, partial decision feedback parameters should increase as the stage continues. Research results in [16, 17] can be introduced to design optimum partial decision feedback parameters.

In Figure 7, the performance comparison of the PMDFE, the MIMO-OFDM [11], the FDLE, and the FDLE-PIC [12] is given for (2Tx, 2Rx). It can be seen that the proposed PMDFE with $p_1 = 0.7$, $p_2 = 0.8$, $p_3 = 0.9$, and $p_4 = 0.95$ has the best performance. If the target BER is $10^{-2}$, the PMDFE can obtain about 4.4 dB gain over the MIMO-OFDM, 2.1 dB gain over the FDLE-PIC, and 3.9 dB gain over the FDLE. As shown in [11], the MIMO-OFDM divides frequency-selective channels into many narrow parallel flat fading channels. At each subchannel, a V-BLAST processor [9] is utilized to detect the desired signal. Therefore, there does not exist multipath diversity or frequency diversity in MIMO-OFDM system. Certainly, MIMO-OFDM scheme with space-time block coding will introduce more spatial diversity to mitigate

| Tap index | Tap delay (ns) | Relative power (dB) |
|-----------|--------------|---------------------|
| 1         | 0            | -3.2                |
| 2         | 100          | -5                  |
| 3         | 400          | -4.5                |
| 4         | 500          | -3.6                |
| 5         | 700          | -3.9                |
| 6         | 800          | 0                   |
| 7         | 950          | -3                  |
| 8         | 1200         | -1.2                |
| 9         | 1350         | -5                  |
| 10        | 1450         | -3.5                |

**Figure 5:** The performance of the PMDFE with partial decision feedback parameters $p_1 = 0.7$, $p_2 = 0.8$, $p_3 = 0.9$, and $p_4 = 0.95$ over (2Tx, 2Rx).

**Figure 6:** The performance comparison between different partial decision feedback parameters over (2Tx, 2Rx).
the fading. However, as proved in [18, 19], space-time block coding will decrease the capacity of MIMO channels. The performance of the FDLE with one-stage PIC is still limited by CCI and ISI. The proposed PMDFE can achieve multipath and spatial diversity simultaneously. Moreover, the great improvement is obtained and the error propagation is reduced when the detection stage continues.

In Figure 8, the performance comparison of the PMDFE, the MIMO-OFDM [11], the FDLE, and the FDLE-PIC in [12] is given for (3Tx, 3Rx). If the target BER is $10^{-2}$, the PMDFE can obtain about 3.2 dB gain over the MIMO-OFDM, 2.5 dB gain over the FDLE-PIC, and 4.8 dB gain over the FDLE.

In Figure 9, the BER performance of the PMDFE with partial decision feedback parameters $p_1 = 0.7$, $p_2 = 0.8$, $p_3 = 0.9$, and $p_4 = 0.95$ over (1Tx, 1Rx), (2Tx, 2Rx), (3Tx, 3Rx), and (4Tx, 4Rx) are compared. When the space-time architecture is reduced to (1Tx, 1Rx), the MISO-DFE is reduced to a SISO-DFE [5]. The performance of the PMDFE will be improved at a high SNR as the number of the transmitter or receiver antennas increases.

7. CONCLUSION

In this paper, a PMDFE is proposed for single-carrier layered space-time systems with the fixed CP. It is equivalent to a DFE in SD combined with a MISO-DFE in FD and TD. It can achieve better performance than the MIMO-OFDM, the FDLE, and the FDLE-PIC.

APPENDIX

For notation simplicity, we define

$$W = W[f], \quad B = B[f], \quad H = (H[f])_r.$$  (A.1)
With the matrix inversion lemma,

\[
(A + ab^H)^{-1} = A^{-1} - \frac{A^{-1}ab^HA^{-1}}{c}, \quad c = 1 + b^HA^{-1}a.
\]

(A.2)

Therefore, we have

\[
\left( \frac{\sigma_d^2}{\sigma_d^2 I} + HH^H \right)^{-1} = \frac{\sigma_c^2}{c} \left[ I - \frac{\sigma_d^2 HH^H}{\sigma_d^2} \right],
\]

(A.3)

where

\[
c = 1 + \frac{\sigma_d^2 \|H\|^2}{\sigma_v^2}.
\]

(A.4)

We compute the first term of the summation in (13). Substituting (15) into (13), we have that

\[
\sigma_d^2 \|W^H H + B - 1\|^2 = \sigma_d^2 \|HH^H \left( \frac{\sigma_d^2}{\sigma_v^2} I \right)^{-1} H(1 - B) + (B - 1) \|^2.
\]

(A.5)

From (A.3), we have

\[
\sigma_d^2 W^H W
\]

\[
= \sigma_d^2 \left[ \frac{\sigma_d^2}{\sigma_v^2} \left( \|H\|^2 - \|H\|^4 \frac{\sigma_d^2}{\sigma_v^2} \right) \right] (1 - B) + (B - 1) \|^2.
\]

(A.6)

With the same derivation, the second term of the summation in (13) can expressed as

\[
\sigma_d^2 W^H W
\]

\[
= \frac{\sigma_d^4}{\sigma_v^2} \left( I - \frac{\sigma_d^2 HH^H}{\sigma_v^2} \right) \left( I - \frac{\sigma_d^2 HH^H}{\sigma_v^2} \right) H^\dagger |1 - B|^2
\]

(A.7)

\[
= \frac{\sigma_d^4}{\sigma_v^2 c^2} |1 - B|^2 \|H\|^2.
\]

With (A.6) and (A.7), we have that

\[
\sigma_d^2 \|W^H H + B - 1\|^2 + \sigma_d^2 W^H W
\]

\[
= \frac{\sigma_d^2 |1 - B|^2}{c} + \frac{\sigma_d^2 |1 - B|^2}{\sigma_v^2 / \sigma_d^2 + \|H\|^2}.
\]

(A.8)

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