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Open loop downlink transmit diversity for TDD: STTD for TDD

Texas Instruments, May 25th, 1999

1.0 Introduction

In the WG1 # 4, the time switched time diversity (TSTD) was accepted as the open loop antenna diversity technique for the downlink of the WCDMA TDD systems. However, TSTD suffers from the power amplifier (PA) imbalance at the base station and results in a higher peak to average ratio (PAR). Also, TSTD does not yield full path diversity and only yields the diversity going through the interleaver. Hence, the TSTD diversity gains are dependent upon the switching patterns and the interleaving scheme chosen. Further, we find that the TSTD raw BER performance is actually slightly worse than the no-diversity (ND) performance when half the users are switched on the first antenna and the other half are switched on the second antenna to maintain PA balance. On the other hand, Space time block coding based transmit antenna diversity (STTD) is a PA balanced scheme and yields full path diversity. Hence, STTD has been accepted as the open loop antenna diversity technique for the WCDMA FDD system. In this document we propose STTD for the WCDMA TDD mode systems. We first derive the combined STTD decoder and the interference canceller, in particular for the zero forcing block linear equalizer (ZF-BLE), for the TDD systems. We denote this as the *exact STTD ZF-BLE*. We then present a *simplified STTD ZF-BLE* whose performance is the same as the exact STTD ZF-BLE, but has lower complexity.

Based upon the raw BER simulations, we find that the E_b/N_0 gain for the *simplified STTD ZF-BLE over no-diversity ZF BLE* systems is between 1.0-3.0 dB and that over the *TSTD ZF-BLE* is slightly higher and is between 1.2-3.5 dB. The total complexity increase of the *simplified STTD ZF-BLE* over the overall complexity of the approximate no-diversity ZF-BLE (*ND ZF*-BLE) [4] is expected to be about 10-15 %, which is not a significant increase.

We thus show that STTD gives substantial performance gains for the TDD system without significantly increasing the complexity of the user equipment (UE).

2.0 Open loop transmit Diversity

2.1 The space time transmit diversity (STTD) scheme for TDD

The basic STTD encoder for TDD is the same as the FDD system [1, 2, figure(8) in 8] and is shown in figure 1 for illustration.

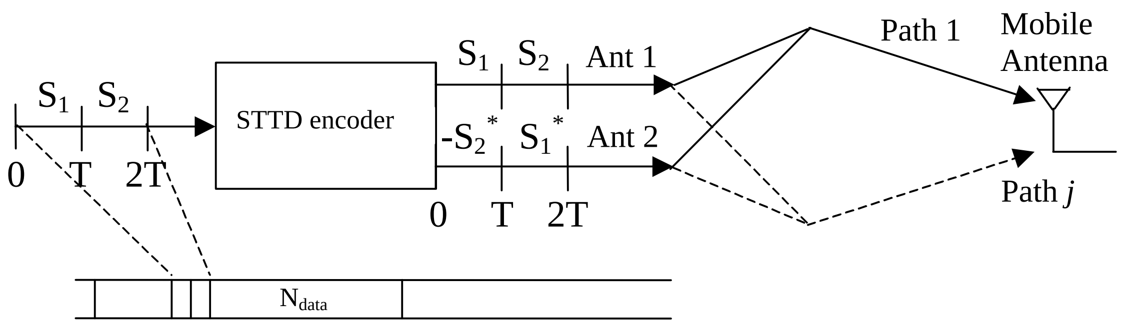

Figure (1): The block diagram of an STTD encoder for data is shown

Hence forth, we follow the same notation as that followed in [3] for analyzing the exact STTD ZF-BLE. Similar to [3], let K be the number of users and M be the number of symbols transmitted, Q the spreading factor and the matrix A denote the composite response of the antenna 1 (including multi-path). Similar to equation (6) in [3], the total received signal for STTD encoded data is now given by;

$$\bar{e} = A\bar{d} + B\bar{d}^* + \bar{n} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots..(1)$$

where $\bar{d}$ is composite transmitted data vector (equation (1) of [3]) and $\bar{d}^*$ is its conjugate. The form of the matrix B is similar to that of matrix A (with a slight rearrangement of elements due to the STTD encoding as shown in figure (1)) and it is the composite channel response for $\bar{d}^*$ transmitted from antenna 2. We now write equation (2) from equation (1) above;

$$\begin{bmatrix} \bar{e} \\ \bar{e}^* \end{bmatrix} = \begin{bmatrix} A & B \\ B^* & A^* \end{bmatrix} \begin{bmatrix} \bar{d} \\ \bar{d}^* \end{bmatrix} + \begin{bmatrix} \bar{n} \\ \bar{n}^* \end{bmatrix} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots..(2)$$

Letting $\begin{bmatrix} \bar{e} \\ \bar{e}^* \end{bmatrix} = \bar{E}$, $\begin{bmatrix} A & B \\ B^* & A^* \end{bmatrix} = C$, $\begin{bmatrix} \bar{d} \\ \bar{d}^* \end{bmatrix} = \bar{D}$ and $\begin{bmatrix} \bar{n} \\ \bar{n}^* \end{bmatrix} = \bar{N}$ we can rewrite the equation (2) as;

$$\bar{E} = C\bar{D} + \bar{N} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots.(3).$$

Because the rows of the top and bottom half of matrix C are linearly independent, the matrix C has full rank. The vector $\bar{E}$ in equation (3) can now be considered to be the net received vector corresponding to the equation (6) in [3]. Following the same procedure as in [3] to derive the ZF-BLE, the corresponding form of equation (25) in [3] is;

$$\hat{\bar{D}}_{ZF-BLE} = (C^H R_N^{-1} C)^{-1} C^H R_N^{-1} \bar{E} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(4)$$

where $C^H = (C^T)^*$. Again letting $R_N = s^2 I_{2KM \times 2KM}$ (the noise correlation matrix) ***we get the exact STTD ZF-BLE to be***;

$$\hat{\bar{D}}_{ZF-BLE} = (C^H C)^{-1} C^H \bar{E} \ldots\ldots\ldots\ldots\ldots\ldots\ldots.(5)$$

Notice that the above equation gives the estimate for both the vectors $\bar{d}$ and $\bar{d}^*$. However, by the construction of equation (2), both the estimate of $\bar{d}$ and the conjugate of the estimate of $\bar{d}^*$ are exactly the same. Hence, it is not essential to obtain the estimates for both $\bar{d}$ and $\bar{d}^*$, and an estimate for only one of them is sufficient.

2.2 Simplified STTD ZF-BLE

In equation (5) of the previous section we have presented the *exact STTD ZF-BLE*. In this section we present the simplified STTD ZF-BLE to reduce the complexity of the exact STTD ZF-BLE. We can see that the main complexity increase for the *exact STTD ZF-BLE* is the matrix inversion for the matrix ($C^H C$) in equation (5). The size of ($C^H C$) is ($2MK \times 2MK$) as against the matrix inversion involved in an *ND ZF-BLE* (equation (25) of [3]) where in the matrix inversion is only of size ($MK \times MK$). Hence, we now look into the details of the matrix inversion of ($C^H C$). Expanding ($C^H C$) we get;

$$C^H C = \begin{bmatrix} A^H & B^T \\ B^H & A^T \end{bmatrix} \begin{bmatrix} A & B \\ B^* & A^* \end{bmatrix} = \begin{bmatrix} A^H A + (B^H B)^* & A^H B + (B^H A)^* \\ (A^H B)^* + (B^H A) & (A^H A)^* + B^H B \end{bmatrix} = \begin{bmatrix} X & Y \\ Y^* & X^* \end{bmatrix} \quad \text{.....(6)}$$

where, $X = A^H A + (B^H B)^*, Y = A^H B + (B^H A)^*$. The principal symbol energy terms are along the main diagonal of the matrix *X*. The off-diagonal terms of the matrix *X* correspond to the ISI and MAI terms from antenna 1 to antenna 1 and antenna 2 to antenna 2 similar to the ND system. The matrix *Y* on the other hand consists of ISI and MAI terms from antenna 1 to antenna 2 and vice versa. Since the principal symbol energies are all concentrated along the main diagonal terms of the matrix *X*, we can see that *det(X)* > *det(Y)* in all cases. In fact for, the ITU channel models indoor, indoor-to-outdoor pedestrian and the vehicular models we find that *independent of the spreading gain Q*, *det(X) >> det(Y).* This allows us to reduce the complexity of the inversion of the matrix $C^H C$. Invoking the matrix inversion lemma [5, A.22] we have,

$$(C^H C)^{-1} = \begin{bmatrix} X & Y \\ Y^* & X^* \end{bmatrix}^{-1} = \begin{bmatrix} X^{-1} + X^{-1} Y \Delta^{-1} Y^* X^{-1} & -X^{-1} Y \Delta^{-1} \\ -\Delta^{-1} Y^* X^{-1} & \Delta^{-1} \end{bmatrix} \quad \text{...............(7)}$$

$$\Delta = X^* - Y^* X^{-1} Y$$

Using the fact that the *det(X) >> det(Y)*, and for a matrix *Z* with abs(det(*Z*)) < 1, $(I - Z)^{-1} = \sum_{i=0}^{\infty} Z^i$ [6] we can approximate the Δ^{-1} in equation (7) as follows;

$$\Delta^{-1} = X^{*\frac{-1}{2}} \left(I - X^{*\frac{-1}{2}} Y^* X^{-1} Y X^{*\frac{-1}{2}} \right)^{-1} X^{*\frac{-1}{2}} \cong X^{*\frac{-1}{2}} \left(I + X^{*\frac{-1}{2}} Y^* X^{-1} Y X^{*\frac{-1}{2}} \right) X^{*\frac{-1}{2}} \quad \text{......(8)}$$
$$= \left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right)$$

Substituting the above approximation back into equation (7) we now get;

$$(C^H C)^{-1} \cong \begin{bmatrix} X^{-1} + X^{-1} Y \left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right) Y^* X^{-1} & -X^{-1} Y \left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right) \\ -\left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right) Y^* X^{-1} & \left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right) \end{bmatrix}$$

..........(9)

We can see that in equation 9, we have to only invert a single matrix *X* of size (*MK x MK*) as against the size (2*MK x 2MK*) inversion required for the matrix ($C^H C$). As mentioned in the end of the section (2.1) we need to estimate only either $\bar{d}$ and $\bar{d}^*$. Defining,

$$F = \left[-\left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right) Y^* X^{-1} \quad \left(X^{*-1} + X^{*-1} Y^* X^{-1} Y X^{*-1} \right) \right] \quad \text{.............(10)}$$

and choosing to obtain $\bar{d}^*$ in equation (5), and using equation (9) we now obtain the ***simplified STTD ZF-BLE*** solution to be;

$$\boxed{\bar{d}^*_{MK \times 1} = F_{MK \times 2MK} C^H_{2MK \times 2MK} \bar{E}_{2MK \times 1}} \quad \text{................................(11)}$$

For the ease of understanding the ***simplified STTD ZF-BLE***, equation (11) explicitly gives the matrix sizes for all the matrices involved.
An mentioned above, the *simplified STTD ZF-BLE* requires the inversion of the matrix *X* of size (*MK x MK*) which the same size as that for regular no-diversity (ND) systems. The matrix *X* for STTD has similar form as the *ND* systems, implying that its inversion complexity can be reduced in the same manner as the approximate *ZF-BLE* [4]. Thus, the complexity of matrix inversion (which is dominant complexity for a ZF-BLE) for STTD is the same as the ND systems. There are extra matrix multiplications as given in equation (10) for the STTD system. However, this does not significantly increase the complexity. For the channel estimation for STTD, an orthogonal preamble will have to be transmitted implying that the channel estimation complexity is doubled. However, since the overall complexity for the *ND ZF-BLE* is dominated by the complexity of the matrix inversion, which is the same for both the ND and STTD systems, we would not expect significant complexity increase for *simplified STTD ZF-BLE* over *ND ZF-BLE*. *Following a procedure similar to [4], we find that the total complexity of the simplified* ***STTD ZF-BLE*** *(including the extra matrix multiplies, channel estimation) should only be about 10-15 % more than the* ***approximate ZF-BLE complexity,*** *as reported in [4] for users K = 1, .., 12. This is not a significant increase, considering the fact that we can achieve diversity gain.*

2.3 Simulations results

We now do link level simulations to evaluate the performance gains of *the simplified STTD ZF-BLE* over the *ND ZF-BLE*. The link level simulations parameters used are given in table 1:

	Vehicular	Indoor-to-outdoor pedestrian
Velocity	120 kmph (Figures 2-4)	3 kmph (Figures 5,6)
Spreading gain (SF)	8, 16	8, 16
Number of users	SF = 8: 4 SF = 16: 4, 8	SF = 8:4 SF = 16: 4
User allocation on antennas	STTD: All users on 1, 2 ND: All users on 1 TSTD: Half users each on 1, 2	STTD: All users on 1, 2 ND: All users on 1 TSTD: Half users each on 1, 2
Channel Estimation	Perfect	Perfect
FEC encoding	No	No
Joint detection	STTD: Yes (simplified STTD ZF-BLE) ND: Yes (ZF-BLE) TSTD: Yes (ZF-BLE)	STTD: Yes (simplified STTD ZF-BLE) ND: Yes (ZF-BLE) TSTD: Yes (ZF-BLE)
Simplified STTD ZF-BLE performance gain over *ND ZF-BLE* (dB)	1.0 dB at raw BER = 10^{-1}	3.0 dB at raw BER = 10^{-2}
Simplified STTD ZF-BLE performance gain over *TSTD ZF-BLE* (dB)	1.2 dB at raw BER = 10^{-1}	3.5 dB at raw BER = 10^{-2}

Table 1: The simulation parameters to compare the performance of STTD against the ND, TSTD system are given. We can see that the performance gains of simplified STTD ZF-BLE over the ND ZF-BLE are between 1.0-3.0 dB and that over TSTD is 1.2-3.5 dB. Because of the time diversity at high Doppler, the raw BER chosen for Vehicular is 10^{-1} *as against the* 10^{-2} *for indoor-to-outdoor pedestrian scenario.*

The performance is shown in figures 2-6:

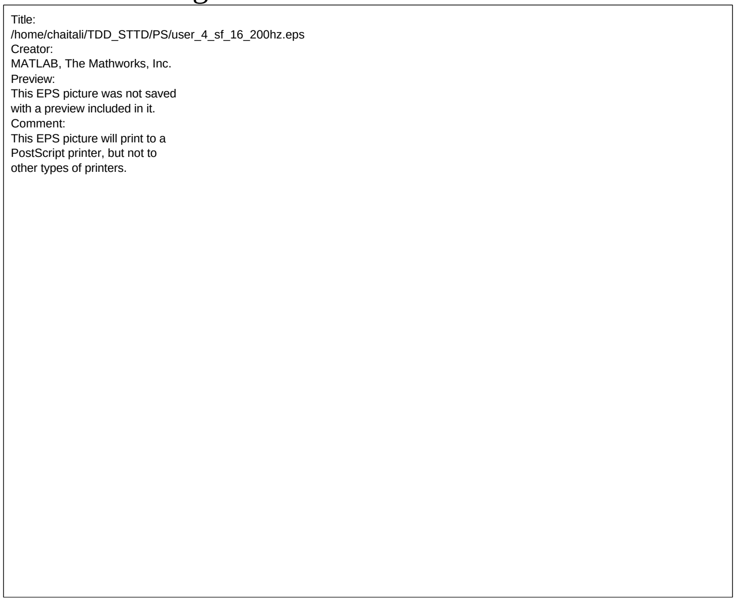

Figure 2: Link level simulations comparing the raw BER performance of simplified STTD ZF-BLE against ND ZF-BLE for spreading gain = 16, number of users K = 4, Vehicular B channel, perfect channel estimates. We can see that STTD gives a diversity gain of about 1 dB at BER = 10^{-1}.

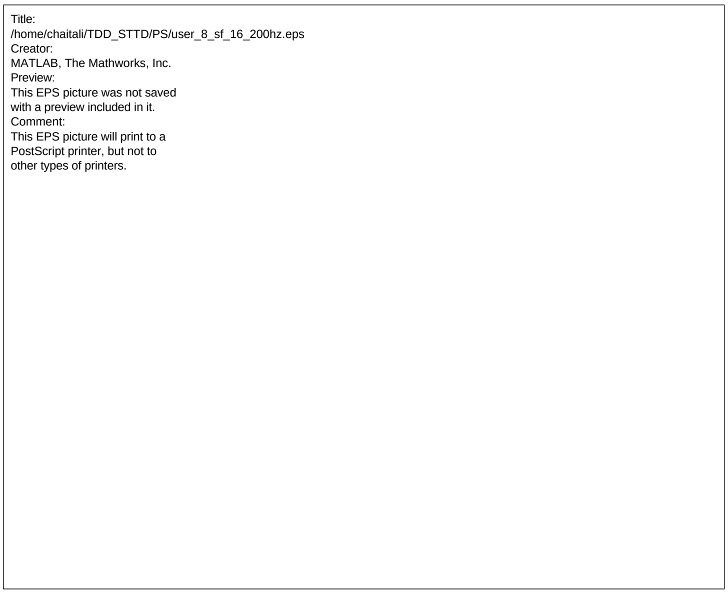

Figure 3: Link level simulations comparing the raw BER performance of simplified STTD ZF-BLE against ND, TSTD ZF-BLE for spreading gain = 16, number of users K = 8, Vehicular B channel, perfect channel estimates. We can see that STTD gives a diversity gain of about 1 dB at BER = 10^{-1} *over ND and 1.2 dB over TSTD. We can also see that the performance of the approximate STTD is the same as the exact STTD implying that the assumption in equation(8) is valid.*

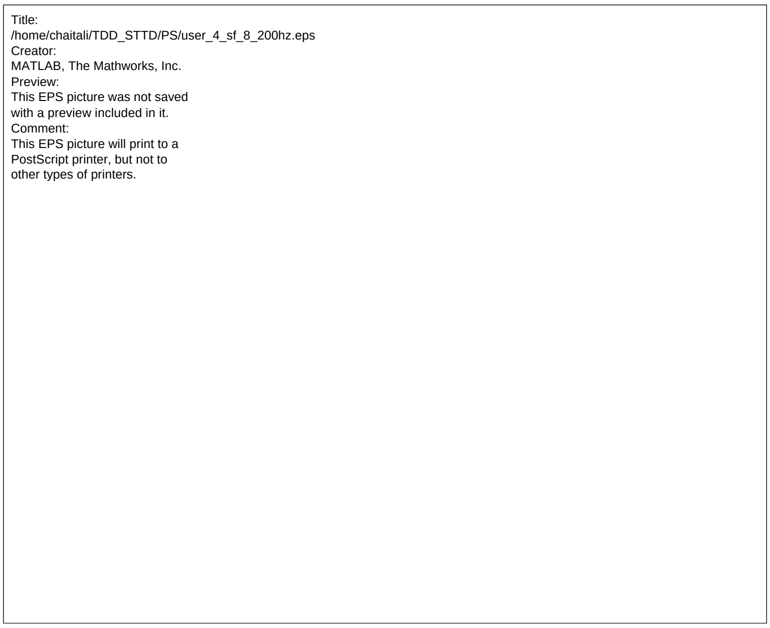

Figure 4: Link level simulations comparing the raw BER performance of simplified STTD ZF-BLE against ND, TSTD ZF-BLE for spreading gain = 8, number of users K = 4, Vehicular B channel, perfect channel estimates. We can see that STTD gives a diversity gain of about 1 dB at BER = 10^{-1} over ND and 1.3 dB over TSTD. We can also see that the performance of the approximate STTD is the same as the exact STTD implying that the assumption in equation(8) is valid.

Figure 5: Link level simulations comparing the raw BER performance of simplified STTD ZF-BLE against ND ZF-BLE for spreading gain = 16, number of users K = 4, Outdoor-to-Indoor Pedestrian channel, perfect channel estimates. We can see that STTD gives a diversity gain of about 3 dB at BER = 10^{-2}.

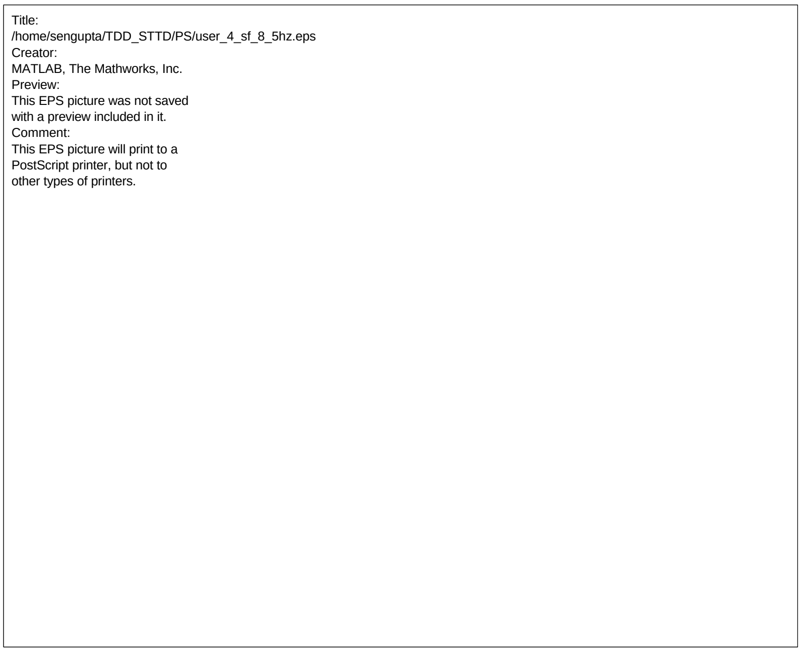

Figure 6: Link level simulations comparing the raw BER performance of simplified STTD ZF-BLE against ND, TSTD ZF-BLE for spreading gain = 8, number of users K = 4, Outdoor-to-Indoor Pedestrian channel, perfect channel estimates. We can see that STTD gives a diversity gain of about 3 dB at BER = 10^{-2} over ND and 3.5 dB over TSTD.

3.0 Conclusions

In terms of raw BER, we have shown that STTD gives significant diversity gains between 1.0-3.0 dB over the ND systems and 1.2-3.5 dB gains over the TSTD without significantly increasing the complexity of the zero forcing block linear equalizer in the user equipment. Based upon the framework given in this paper, the minimum mean squared error (MMSE) detector combined with the STTD decoding can be derived in a similar manner and it can be shown that its complexity increase over the MMSE for ND is also not significantly higher. Thus, we can conclude that using STTD for TDD systems will not significantly increase the complexity of the joint detection at the mobile.

From figures 3, 4 and 6 we can see that TSTD actually does slightly worse than ND (in terms of raw BER). The reason for this happening is that half of the interference comes from the transmission on the second antenna to maintain PA balance, making the simulations slightly different from the ND simulation. The ZF-BLE does not do a perfect interference cancellation. The net effect is that the TSTD raw BER is worse than the ND raw BER. Hence the STTD gains over TSTD in terms of raw BER are nominally more than the gains over ND. Further, STTD has the advantage of being a power-balanced scheme, implying that the peak to average ratio (PAR) of the base station power amplifier is in general lower than the TSTD scheme. Also, STTD gives the full path diversity independent of the switching patterns, as against the TSTD, whose diversity gains will depend upon the interleaving schemes and the switching pattern [7]. Similarly, STTD provides path diversity independent of the number of switching points, as against the

closed loop diversity techniques [7], which require multiple switching points to reduce performance degradation.
Hence STTD's advantage in terms of providing better path diversity gains without significant increase in mobile complexity, makes it a better choice for open loop antenna diversity.

References

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