Signal Detection for Distributed Space-Time Block Coding: 4 Relay Nodes under Quasi-Synchronisation

F.-C. Zheng, A. G. Burr, and S. Olafsson

Abstract—Most research on Distributed Space-Time Block Coding (D-STBC) has so far focused on the case of 2 relay nodes and assumed that the relay nodes are perfectly synchronised at the symbol level. This paper applies STBC to 4-relay-node systems under quasi-synchronisation and derives a new detector based on parallel interference cancellation, which proves to be very effective in suppressing the impact of imperfect synchronisation.

Index Terms—Distributed space-time block coding, distributed transmit diversity, cooperative diversity, and cooperative relay.

I. INTRODUCTION

There has been a considerable research effort on creating and harnessing space diversity by applying Space-time block coding (STBC) [1] in a distributed fashion (termed distributed STBC or D-STBC) to single antenna systems [2]-[10]. So far, however, most research on D-STBC has (i) focused on the case of 2 relay nodes (using 2-Tx STBC, i.e. the now famed Alamouti scheme) and (ii) assumed that the relay nodes are perfectly synchronised at the symbol level. In practice, however, the assumption of accurate symbol-level synchronisation cannot be extended to the case of more than two relay nodes. As such, some pioneering work has recently been carried out on D-STBC for 2 relay node systems under imperfect synchronisation (e.g. [11] and [12]), chiefly using sequence detection or equalisation techniques at the destination node. Compared with the original STBC schemes, however, these existing methods may incur a much higher computational complexity at the receiver. To address this issue, a near-optimum detector yet with near-Alamouti simplicity is presented in [14] and [15]. The detector in [14] and [15], however, cannot be extended to the case of more than two relay nodes.

Considering the higher diversity order (DV) that a system with more than 2 relay nodes can offer, a logical question to ask is therefore: how would a 3 or 4-relay-node STBC system perform under imperfect synchronisation? To our best knowledge, little has so far been reported on this important issue. To this end, this paper applies STBC to 4-relay-node systems under rough or quasi-synchronisation and derive the corresponding detectors (with the 3 relay node systems being a special case).

It is shown that, similarly to the 2-relay node case, imperfect synchronisation in the 4-relay-node case also causes significant performance degradation to the conventional detector. To combat such a degradation, we propose a new STBC detector based on quasi-synchronisation and the principle of parallel interference cancellation (PIC). The PIC detector is moderate in computational complexity, but proves to be very effective in suppressing the impact of imperfect synchronisation. To further increase the diversity order (DV), maximum ratio combining (MRC) is used to soft-combine the signal received at the broadcasting phase and that received at the relay phase. These form the major contributions of this paper.

For the rest of this paper, [.]T and [.]H represent “transpose” and “conjugate”, respectively. CN(0, σ2) denotes the set of Gaussian distributed complex numbers with the standard variance of σ2 (i.e. 0.5σ2 per dimension).

II. D-STBC WITH IMPERFECT SYNCHRONISATION

This paper deals with the 4-relay node model depicted in Fig. 1, comprising the source node (S), the destination node (D), and 4 relay nodes (Rm, m = 1, ..., 4). Obviously, a system with 3 relay nodes is merely a special case of Fig. 1 where h_d (the channel gain from R4 to D) is zero. As in almost all cooperative relay systems [4]-[6], there are two phases involved: Phase 1 for broadcasting and Phase 2 for relaying.

[Phase 1] Node S transmits while nodes R_m’s and D receive. To prepare for the STBC operation in Phase 2, the symbols transmitted by S are grouped into 3-symbol tuples. Denoting the ith such tuple as s(i) = [s(1, i), s(2, i), s(3, i)]T, the corresponding signal received at D after this direct transmission (DT) is then

\[ r_{sd}(i) = h_{sd} s(i) + n_{sd}(i), \]

where h_{sd} is the channel gain between S and D, r_{sd}(i) is \([r_{sd}(1, i), r_{sd}(2, i), r_{sd}(3, i)]^T\), n_{sd}(i) = \([n_{sd}(1, i), n_{sd}(2, i), n_{sd}(2, i)]^T\), and n_{sd}(l, i) \in CN(0, \sigma_n^2) is the additive noise. The maximum likelihood detection can be carried out via least square (LS) search:

\[ \hat{s}_{sd}(l, i) = \arg \min_{s_j \in S} |h_{sd}^* r_{sd}(l, i) - |h_{sd}|^2 s_j|^2 \] (1a)

with l = 1, 2, 3. S is the alphabet containing M entries or symbols for M-QAM or M-PSK.

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accurate synchronisation is difficult or impossible [11]-[15],

there is normally a timing misalignment of $\tau_m$ between the received versions of these signals, as is shown in Fig. 2.

As a rough synchronisation is always required, we assume a condition of quasi-synchronisation in this paper: $\tau_m \in [0, T_q]$, where $T_q < T$ (the symbol period). This assumption is much easier to meet in practice (see Section IV). Even for a flat fading channel which this paper assumes (i.e. the one-to-one channel itself for any transmit/receive pair is flat fading), such a relative time delay (Fig. 2) will still cause “inter symbol interference (ISI)” from neighbouring symbols at the receiver, due to sampling or matched filtering (whatever kind of pulse shaping is used).

Let $h_m(l)$, with $l = 0, 1, 2, \ldots, \infty$, be the channel gain between $R_m$ and $D$ under quasi-synchronisation and $n_{rd} \in \mathbb{C}N(0, \sigma_r^2)$ be the additive noise. Here, $h_m(l)$ represents the composite effect of time delay $\tau_m$ and the pulse shaping waveform (PSW) and reflects the inter symbol interference (ISI) from the previous symbols (due to imperfect synchronisation). For practical PSW’s such as the raised cosine, $h_m(2)$ is already much less dominant. All the other terms such as $h_m(3)$ are even smaller or zero in value (e.g. due to digital generation of the PSW) and can therefore be truncated (otherwise they can be dealt with similarly using the procedure in this paper). With no loss of generality, we can assume that the receiver at $D$ is perfectly synchronised to $R_1$, i.e., $\tau_1 = 0$ and $h_1(0) = h_1$. The received signal at $D$ over $4T$ can then be represented by

$$r(1,i) = \sum_{m=1}^{4} h_m(0) x_m(1,i) + \sum_{m=2}^{4} h_m(1) x_m(4,i-1) + \sum_{m=2}^{4} h_m(2) x_m(3,i-1) + n_{rd}(1,i), \quad \text{ (3a)}$$

$$r(2,i) = \sum_{m=1}^{4} h_m(0) x_m(2,i) + \sum_{m=2}^{4} h_m(1) x_m(1,i) + \sum_{m=2}^{4} h_m(2) x_m(4,i-1) + n_{rd}(2,i), \quad \text{ (3b)}$$

$$r(k,i) = \sum_{m=1}^{4} h_m(0) x_m(k,i) + \sum_{m=2}^{4} h_m(1) x_m(k-1,i) + \sum_{m=2}^{4} h_m(2) x_m(k-2,i) + n_{rd}(k,i), \quad k = 3, 4. \quad \text{ (3c)}$$

Fig. 2. The time delay (under quasi-synchronisation) at Node $D$: the received signal at $D$ is a superposition of 4 symbol streams: those from $R_1$, ..., $R_4$.

If Phase 1 is successful, then there is no need to solicit the help of any relay nodes. Otherwise, the following phase will follow. This is termed “incremental relay” in [4] (Namely, “relay” is invoked only if necessary (e.g. when Phase 1 fails) and therefore on an incremental basis). Even if Phase 1 fails, however, (1) still contains valuable information and should be fully utilised (as is shown later in this paper).

[Phase 2] The symbols detected at $R_1$ are encoded using the 4-Tx STBC structure and transmitted to $D$. Due to its simpler form, the square 4-Tx STBC structure in Eq. (3.39) of [16] is used here (compared with the $\mathcal{H}_4$ code in [1]). Obviously, the relay mode of “decode and forward” is employed in this paper. To this end, a sufficient level of cyclic redundancy check (CRC) can be incorporated into the data packet at $S$ so that relaying will happen only if the packet is correctly detected at $R_m$’s (i.e. “selective relay” [4]). Also in Fig. 1, $h_m$ is the channel gain from $R_m$ to $D$ under perfect inter-relay-node synchronisation.

Denoting the encoded symbol stream at $R_m$ corresponding to $s(i)$ as $x_m(i) = [x_m(1,i), x_m(2,i), x_m(3,i), x_m(4,i)]^T$, the STBC matrix [16] transmitted by $R_m$’s is therefore

$$\begin{bmatrix} x_1(i), & x_2(i), & x_3(i), & x_4(i) \end{bmatrix} = \begin{bmatrix} s(1,i) & s(2,i) & s(3,i) & 0 \\ -s^*(2,i) & s^*(1,i) & 0 & -s(3,i) \\ -s^*(3,i) & 0 & s^*(1,i) & s(2,i) \\ 0 & s^*(3,i) & -s^*(2,i) & s(1,i) \end{bmatrix}, \quad (2)$$

meaning that Node $R_m$ transmits $x_m(k,i)$ for the $k$th symbol period corresponding to the $i$th symbol tuple.

Due to factors such as different propagation delays, $x_m(i)$’s will most likely arrive at $D$ at different time instants. Since accurate synchronisation is difficult or impossible [11]-[15],
The relative strength of \( h_m(l), l = 1, 2, \) will be represented by ratio
\[
\beta_m(l) = \frac{|h_m(l)|^2}{|h_m(0)|^2}
\]
(3d) in this paper. The value of \( \beta_m(l) \) depends upon delay \( \tau_m \) (or timing mismatch) and the particular pulse shaping waveform used. However, we always have \( \beta_m(0) = 0 \) for \( \tau_m = 0 \), and \( \beta_m(1) = 1 \) (i.e. 0 dB) for \( \tau_m = 0.5T \) (due to the symmetry of PSW’s).

All the channel gains above are assumed to be block Rayleigh fading (from data packet to packet), i.e., \( h_{sd} \in \mathbb{CN}(0, \sigma^2_d) \), and \( h_m \in \mathbb{CN}(0, \sigma^2_m) \). For a fair comparison with the non-relay schemes, \( R_m \) transmits at 1/4 power, i.e. \( \sigma^2_r = \frac{\sigma^2_d}{4} \).

Substituting (2) into (3a-c), we have
\[
\mathbf{r}(i) = \mathbf{H}_3 \mathbf{s}(i) + \mathbf{B} \mathbf{s}^*(i) + \mathbf{I}(i) + \mathbf{n}_r, \quad (4)
\]
where
\[
\mathbf{r}(i) = [r(1, i), r^*(2, i), r^*(3, i), r^*(4, i)]^T,
\]
\[
\mathbf{H}_3 = \begin{bmatrix}
    h_1(0) & h_2(0) & h_3(0) \\
    h_2(0) & -h_1^*(0) & 0 \\
    h_3(0) & 0 & -h_2^*(0) \\
    0 & -h_3^*(0) & h_2(0)
\end{bmatrix},
\]
\[
\mathbf{B} = \begin{bmatrix}
    h_2(0) & 0 & 0 \\
    0 & h_3(0) & 0 \\
    h_3(0) & 0 & 0
\end{bmatrix},
\]
\[
\mathbf{I}(i) = [I(1, i), I(2, i), I(3, i), I(4, i)]^T
\]
and
\[
\mathbf{n}_r(i) = [n_r(1, i), n_r(2, i), n_r(3, i), n_r(4, i)]^T.
\]

Also,
\[
I(1, i) = \sum_{m=2}^{4} h_m(1) x_m(4, i-1) + \sum_{m=2}^{4} h_m(2) x_m(3, i-1),
\]
(5a)
\[
I^*(2, i) = E^*(2, i) + E^*(2, i-1)
\]
(5b)
with
\[
E^*(2, i) = \sum_{m=2}^{4} h_m(1) x_m(1, i),
\]
\[
E^*(2, i-1) = \sum_{m=2}^{4} h_m(2) x_m(4, i-1),
\]
and
\[
I^*(k, i) = \sum_{m=2}^{4} h_m(1) x_m(k-1, i) + \sum_{m=2}^{4} h_m(2) x_m(k-2, i),
\]
(5c)
\[
k = 3, 4.
\]

Here, \( x_m(k, i-1) \) is from \( \mathbf{s}(i-1) \) according to (2). Note that for the 3 relay node case, \( h_4(l) = 0 \) (hence \( \mathbf{B} = \mathbf{0} \)).

To carry out the detection, we now re-write (4) into the following “real number” form (after some straightforward matrix manipulations):
\[
\begin{bmatrix}
\mathbf{r}_R \\
\mathbf{r}_I
\end{bmatrix} = \begin{bmatrix}
\mathbf{H}_R & \mathbf{s}_R & \mathbf{I}_R & \mathbf{n}_{r,R} \\
\mathbf{H}_I & \mathbf{s}_I & \mathbf{I}_I & \mathbf{n}_{r,I}
\end{bmatrix},
\]
(6)
where \( \mathbf{H}_R = [\mathbf{H}_{3R} + \mathbf{B}_R, -\mathbf{H}_{3I} + \mathbf{B}_I] \), and \( \mathbf{H}_I = [\mathbf{H}_{3I} + \mathbf{B}_I, \mathbf{H}_{3R} - \mathbf{B}_R] \).

Here, for any matrix or vector \( z \), we denote \( z_R = \text{real}(z) \), and \( z_I = \text{imag}(z) \). For brevity, we have dropped the time index \( i \).

It can be shown that matrix \( \mathbf{Q} = \begin{bmatrix} \mathbf{H}_R & \mathbf{H}_I \end{bmatrix} \) in (6) is orthogonal and \( \mathbf{A} = \mathbf{Q}^H \mathbf{Q} = \text{diag}(\lambda, \ldots, \lambda) \), where \( \lambda = \sum_{k=1}^{4} |h_k(0)|^2 \). As such, applying linear transform \( \mathbf{\Theta} = \mathbf{Q}^H \) to (6) does not change the independence property of the additive noise:
\[
\mathbf{g}(i) = \mathbf{\Theta} \begin{bmatrix} \mathbf{r}_R \\
\mathbf{r}_I \end{bmatrix} = \begin{bmatrix} \mathbf{s}_R \\
\mathbf{s}_I \end{bmatrix} + \begin{bmatrix} \mathbf{I}_R \\
\mathbf{I}_I \end{bmatrix} + \mathbf{v},
\]
(7)
where \( \mathbf{g}(i) = [g(1, i), \ldots, g(6, i)]^T \), and \( \mathbf{v} = \begin{bmatrix} \mathbf{n}_{r,R} \\
\mathbf{n}_{r,I} \end{bmatrix} \) is the post-transform additive noise. The corresponding complex number form for \( \mathbf{g}(i) \) can then be constructed:
\[
\mathbf{g}_c(i) = \begin{bmatrix} g(1, i) + j g(4, i) \\
g(2, i) + j g(5, i) \\
g(3, i) + j g(6, i) \end{bmatrix}, \quad j = \sqrt{-1}.
\]
(8)

As was mentioned earlier, even if the DT in Phase 1 fails, Equation (1) still contains valuable information and should therefore be combined with the signal received at Phase 2. As such, we now apply maximum ratio combining (MRC) to (1) and (4), and this immediately leads to the following signal for detection:
\[
\mathbf{f}(i) = [f(1, i), f(2, i), f(3, i)]^T = h_{sd} \mathbf{r}_{sd}(i) + \mathbf{g}_c(i).
\]
(9)

It is easy to show that \( \mathbf{v}_f(i) \), the additive noise vector in (9), still satisfies the property of noise independence below:
\[
\mathbf{E}(\mathbf{v}_f(i) \mathbf{v}_f^H(i)) = \sigma^2_n \mathbf{A}_f,
\]
(10)
where \( \mathbf{A}_f = \text{diag}(\lambda_f, \lambda_f, \lambda_f, \lambda_f, \lambda_f) \), and \( \lambda_f = |h_{sd}|^2 + \lambda = |h_{sd}|^2 + \sum_{k=1}^{4} |h_k(0)|^2 \).

The conventional STBC detector (which assumes “perfect synchronisation”) is then equivalent to the following Least Square (LS) process:
\[
\hat{s}(l, i) = \arg\min_{\mathbf{s} \in \mathcal{S}} |f(l, i) - \lambda_f s_l|^2, \quad l = 1, 2, 3,
\]
(11)
where again \( \mathcal{S} \) is the alphabet containing \( M \) symbols for MQAM or M-PSK. The expression for \( \lambda_f \) indicates that, with 4 relay nodes, a diversity order (DV) of 5 can be achieved (or a DV of 4 if there are only 3 relay nodes).

Due to the interference component of \( \mathbf{\Theta} \begin{bmatrix} \mathbf{I}_R \\
\mathbf{I}_I \end{bmatrix} \) in (6), however, the above procedure can suffer from significant detection errors, unless \( h_{m}(l) = 0 \) for \( l = 1, 2 \) (i.e. \( \tau_m = 0 \), the case of perfect synchronisation). This causes the performance of the conventional STBC detector to be very sensitive to imperfect synchronisation.

III. PIC BASED DETECTION

To mitigate the above impact of imperfect synchronisation, clearly we must remove as much as possible the interference component of \( \mathbf{I}(i) \) in (4) or (6). To this end, we now apply the principle of parallel interference cancellation (PIC), assuming perfect channel state information at D. PIC is a well known
scheme in CDMA systems and is very effective in suppressing multiuser interference [17].

Upon examining (5a) and (5b), we notice that \( x_m(l, i - 1) \), i.e. \( s(i - 1) \), is in fact already known if the detection process has been started properly (e.g. through the use of pilot symbol(s) at the start of the packet). As such, ISI components \( I(1, i) \) in (5a) and \( E(2, i - 1) \) in (5b) can be removed before the PIC iteration. The PIC iteration process can then be conducted as follows.

**[Initialisation]** Set iteration number \( k = 0 \).
- Remove ISI : 
  \[
  r'(1, i) = r(1, i) - I(1, i), \quad \text{and} \quad r'(2, i) = r(2, i) - E^*(2, i - 1).
  \]
- Due to the poor performance of the conventional detector, we now use the detection result of DT (i.e. Eq. (1a)) to initialise \( s(k) \) : 
  \[
  s^{(0)}(i) = [s^{(0)}(1, i), s^{(0)}(2, i), s^{(0)}(3, i)]^T = [s_{sd}(1, i), s_{sd}(2, i), s_{sd}(3, i)]^T.
  \]

**[Iteration]** \( k = 1, 2, \ldots, K \)
- Remove more ISI by calculating
  \[
  r^{(k)}(i) = \begin{bmatrix}
  r'(1, i) \\
  r'(2, i) - E^{(k-1)}(2, i) \\
  r'(3, i) - I^{(k-1)}(3, i) \\
  r'(4, i) - I^{(k-1)}(4, i)
  \end{bmatrix},
  \]
  where \( E^{(k-1)}(2, i) \) and \( I^{(k-1)}(m, i) \) are determined using \( s^{(k-1)}(i) \) via \( x_m^{(k-1)}(l, i) \) in (5a) and (5b).
- Obtain \( g^{(k)}(i) = \Theta \left[ \begin{array}{c}
  \text{real}(r^{(k)}(i)) \\
  \text{imag}(r^{(k)}(i))
  \end{array} \right] \) (i.e. the left hand side of (6)), and its corresponding complex number form:
  \[
  s^{(k)} = \begin{bmatrix}
  g^{(k)}(1, i) + j g^{(k)}(4, i) \\
  g^{(k)}(2, i) + j g^{(k)}(5, i) \\
  g^{(k)}(3, i) + j g^{(k)}(6, i)
  \end{bmatrix},
  \]
- Carry out MRC by calculating
  \[
  f^{(k)}(i) = h_{sd}^\ast f_{sd}(i) + s^{(k)}(i).
  \]
- Apply the LS detection to \( f^{(k)}(i) \) by replacing \( f(i) \) in (11) with \( f^{(k)}(i) \). The detection result is 
  \[
  s^{(k)}(i) = [s^{(k)}(1, i), s^{(k)}(2, i), s^{(k)}(3, i)]^T.
  \]

If the interference \( I(i) \) in (4) or (6) is removed completely, the above PIC detector would achieve maximum likelihood due to the STBC structure and (10). In general, however, it is a sub-optimum scheme. Nonetheless, the PIC detector is still very effective in error reduction and normally only 2 - 3 iterations will deliver most of the performance gain, as will be shown in Section V.

**IV. OTHER ISSUES**

**[Computational Complexity]** For each extra iteration, only (11)-(13) need to be carried out again. The extra computation involved is therefore approximately \( K(9 + M) \) complex multiplications per symbol for \( K \) iterations, where \( M \) is the size of the symbol alphabet. As normally \( K = 2 \) or 3, the increase in computational complexity is still moderate, especially considering the significant performance improvement.

**[Impact of Error Propagation (EP)]** The above scheme relies on the detection result for the previous symbol tuple, therefore can be viewed as a tuple based “decision feedback” method. Naturally, any error in the feedback will have a negative impact on the detection error for the current symbol tuple. Fortunately, such an impact of error propagation (EP) is normally minor, as is shown in the next section.

**[Rough or Quasi- Synchronisation]** This paper assumes \( \tau_m \in [0, T_q] \), where \( T_q < T \) (the symbol period). Otherwise, similarly to the case of asynchronous CDMA systems, the time advanced node can simply be instructed once to delay its transmission by \( kT_q \) if \( kT_q < \tau_m < (k + 1)T_q \) via a feedback channel (\( k = 0, 1, 2, \ldots \)). Such comparison and feedback operations can be carried out during the cooperation setup and channel estimation stage. Note that the feedback here is an easy-to-quantise threshold (e.g. 0.5) not the actual delay or misalignment, and is therefore is much easier to realise in practice. On the other hand, the case of accurate synchronisation corresponds to \( T_q = 0 \), which either is hard to achieve (e.g. due to synchronisation accuracy) or involves much more signaling overhead (e.g. due to more accurate and/or frequent feedback in case of channel variation).

**V. SIMULATION EXAMPLES**

With an 8-PSK system and assuming perfect detection at \( R_m \)’s, the square 4-node D-STBC in (2) is employed (with 4 relay nodes). The signal to noise ratio (SNR) is defined as \( \text{SNR} = \frac{\sigma_r^2}{\sigma_n^2} \) (dB), and all \( R_m \)’s transmit at 1/4 power.

**A. Number of PIC iterations**

Considering that \( h_m(l) \) represents the composite effect of both time delay \( \tau_m \) and the pulse shaping waveform (PSW) and \( h_m(1) \) is normally the most dominant term, we assume in this example that the ratio \( \beta_m(1) = 0.5 \) (i.e. -3 dB) and \( \beta_m(2) = 0 \). The bit error rates (BER’s) of the PIC scheme for \( K = 1, 2, \) and 3 iterations are plotted in Fig. 3. For comparison, also included in the figure are: the corresponding results of the conventional detector here performs even worse than the DT, (ii) the PIC scheme is very effective to mitigate the impact of imperfect or quasi-synchronisation, and (iii) 2 or 3 iterations deliver almost all the gain.

**B. Level of inter-symbol interference (ISI)**

With the same setting as the above, the level of ISI is now changed by varying the value of \( \beta_m(1) \). The BER results of the PIC scheme (\( K = 3 \) iterations) under \( \beta_m(1) \in [-6, -3, 0, \) and 3 (dB) are presented in Fig. 4. For comparison, the corresponding results of the conventional detector are shown in Fig. 5. Again, the BER’s for DT alone and STBC with perfect synchronisation are included as references. It can be seen that the PIC detector is very effective under quasi-synchronisation while the conventional detector has failed in all cases.
c. Practical pulse shaping waveform (PSW) and impact of error propagation (EP)

In this example, the practical PSW of raised cosine with a roll-off factor of 0.22 (e.g. 3GPP) is employed. The values of \( h_m(l) \) therefore now depend upon \( \tau_m \) only. Two scenarios are studied: (i) with error propagation (EP, i.e. the detector is run as is and the detected symbols for the previous tuple are used for the detection of the current tuple), and (ii) with no EP (i.e. the true symbols in the previous tuple are used for the detection of the current tuple). The BER results for \( \tau_m = 0.25T \) and \( 0.5T \) are shown in Fig. 6. Clearly, the PIC detector is still very effective under this practical PSW and the impact of feedback error is almost negligible.

In all the above examples, the PIC detector performed very well even for some large delay values while the conventional detector simply failed, indicating that the proposed PIC detector requires a quasi-synchronisation only (e.g. within \( T_q = 0.5T \)).

Finally, although not presented here due to the space limit, our work has shown that the above PIC procedure can also be implemented without MRC and still deliver significant performance gain (which obviously will be less than that for the case of using MRC).

VI. CONCLUSIONS

D-STBC has been applied to the systems of 3 or 4 relay nodes with the MRC of Phases 1 and 2. The impact of imperfect synchronisation was examined, and a PIC based detection scheme has been proposed. It has been shown that, even with MRC, the conventional STBC detector is very sensitive to imperfect synchronisation, while the PIC based detector, with moderate computational complexity, is very effective in suppressing the impact of even some large timing mismatches (as long as they are within the quasi-synchronisation range).
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