# On the Performance of Turbo Codes-Based Hybrid ARQ with Segment Selective Repeat in WCDMA

Tao Shi and Lei Cao

*Abstract:* **In this paper, a new turbo codes-based hybrid automatic repeat request (TC-HARQ) scheme with segment selective repeat (SSR) is proposed. The main strategy is, upon retransmission, to repeat the data that are most important for the next round of decoding based on the distribution of residual errors after current decoding. The performance in terms of reliability and throughput is analyzed. To adapt to correlated fading channels where an interleaver is always employed before transmission, we further modify the SSR strategy so that data having experienced correlated deep fading are selected for retransmission. Finally, this proposed scheme is applied to the wideband code division multiple access (WCDMA) system under frequency selective fading channels. Simulation results demonstrate that in all single and multiple user cases, SSR-based TC-HARQ leads to significant throughput improvement with similar bit error rate (BER) performance as compared to type-I TC-HARQ.**

*Index Terms:* **Hybrid automatic repeat request (HARQ), segment selective repeat, turbo codes, WCDMA.**

## I. INTRODUCTION

The development of the third generation (3G) mobile communication systems facilitates the transmission of data traffic in wireless environment. However, the high rate data transmission is vulnerable under mobile hostile channels. Although turbo coding [1] has opened a new chapter in forward error correction (FEC), residual errors may still exist after decoding, which stimulates the study of turbo codes-based hybrid automatic repeat request (TC-HARQ). Type-I and type-II TC-HARQ have been proposed in [2] and [3], respectively. However, neither of these strategies considers the distribution of residual errors after decoding, and their retransmission strategies are predetermined before transmission. To retransmit data adaptively, Shea proposed a reliability-based method by utilizing the soft output information of turbo decoding [4]. While this scheme identifies the most "wanted" data in retransmission, it also requires a large amount of feedback information to index those retransmitted bits. This is unpractical because the feedback channel for HARQ is of low data rate in general. With a high data rate, the feedback will occupy much bandwidth of the system's reverse link and is not error free.

In order to generate efficient feedback and retransmission, the segment selective repeat (SSR) concept was proposed by the authors [5], where retransmission is based on data segments whose residual error status are estimated through soft values after decoding. Simulations showed that SSR has significant improvement in throughput efficiency over type-I TC-HARQ under additive white Gaussian noise (AWGN) and Rayleigh fading channels. In this paper, we first present the idea of SSR and further analyze its performance in terms of reliability and throughput. For a practical communication system, where an interleaver is always employed after turbo encoding to counteract the correlated fading channel, we then propose a modified SSR strategy. The basic idea is to find and retransmit the data that experience deep fading together. Finally, we apply the modified SSR-based TC-HARQ to the WCDMA system under frequency selective fading channels with single and multiple user cases. Simulation results demonstrate that it achieves much higher throughput than type-I TC-HARQ but with little compromise in BER.

The rest of the paper is organized as follows. Section II presents the principle of SSR and its performance analysis. The modified SSR is proposed in Section III. Section IV describes the WCDMA environment and channel conditions for implementing SSR-based TC-HARQ. Simulation results comparing the performance of SSR with type-I TC-HARQ are provided in Section V. Section VI concludes the paper and discusses possible future research.

## II. SSR-BASED TC-HARQ

### A. SSR-Based TC-HARQ and Residual Error Status Estimation

Traditional TC-HARQ schemes retransmit either the entire frame [2] (type-I TC-HARQ), or partial frame of incremental redundancy [3] (type-II TC-HARQ). The retransmission strategies are pre-defined and do not adjust adaptively with channel conditions in the sense that the distribution of residual errors in received frames after decoding is not considered in retransmission. As a result, the retransmitted data are equally important for the decoding of each individual bit in the next round of decoding. Consequently, the retransmitted data may be simultaneously superfluous for the data that had already been correctly decoded and inadequate for the data that had been severely corrupted by the channel.

Actually, after turbo decoding, though uniformly distributed residual errors do occur in high error-floor or non-convergence situations, which can often be handled by sophistically designing the encoder and interleaver, it was shown that residual errors are likely to appear in clusters [6]. This suggests that it would be more efficient to only retransmit the data that are highly associated with the error bits. These data are very important for the next round of turbo decoding but have relatively small data size compared with the entire frame. Therefore the throughput, defined as the ratio between the number of all accepted bits at the receiver and the number of all transmitted bits from the transmitter, could be significantly improved with similar bit error rate

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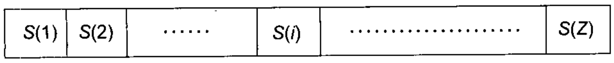


Fig. 1. Segmentation of a frame in SSR.

(BER) as compared to type-I TC-HARQ. In our proposed SSR-based TC-HARQ, each frame is uniformly divided into $Z$ segments with equal size, as shown in Fig. 1. Upon decoding errors detected, all segments are estimated for residual error status, and only a few segments estimated with the most severe corruption will be retransmitted.

In order to estimate the residual error status of different segments after turbo decoding, we propose to use the soft output information after turbo decoding. Turbo codes employ iterative decoding, where the performance is generally improved in each iteration with the aid of a priori information, which is the extrinsic information generated by the other component decoder [1]. Consider the $i$-th iteration of turbo decoding, let $L_m^{(i)}(x)$ and $L_{e_m}^{(i)}(x)$ denote the log-likelihood ratio (LLR) and the extrinsic information of bit $x$ by the $m$-th component decoder ($m = 1, 2$), respectively. It is known [7] that

$$L_1^{(i)}(x) = L_{e_2}^{(i-1)}(x) + \frac{2}{\sigma^2}y_x + L_{e_1}^{(i)}(x) \quad (1)$$

$$L_2^{(i)}(x) = L_{e_1}^{(i)}(x) + \frac{2}{\sigma^2}y_x + L_{e_2}^{(i)}(x) \quad (2)$$

where $(2/\sigma^2)y_x$ is the channel soft output value of bit $x$.

The sign of $L_2^{(i)}(x)$ at the last decoding iteration determines the value of $x$, and its magnitude accounts for the probability that bit $x$ is correctly decoded, i.e., a larger magnitude denotes a higher probability of correct decoding. It was also shown [8] that the mean absolute LLR value over a frame is inversely proportional to the noise power (i.e., $\sigma^2 \propto 1/E[|L_2(x)|]$). Hence, a larger mean absolute LLR value denotes less noise power and therefore less error after decoding. In this research, we adopt the mean absolute LLR value to estimate the residual error status of each data segment. Suppose one frame is uniformly divided into $Z$ segments, each with $l$ bits. We define

$$\overline{|LLR|}_z = \frac{1}{l}\sum_{j=1}^{l} |L_2(X((z-1)l+j))| \quad (3)$$

as the mean absolute LLR value of the $z$-th segment, where $z = 1, 2, \cdots, Z$, $X$ is the systematic bit sequence, and $L_2(X((z-1)l+j))$ denotes the LLR of the $j$-th bit in the $z$-th segment from the 2nd component decoder after the final decoding iteration. In our SSR-based TC-HARQ using mean absolute LLR value, the entire frame is transmitted at the 1st transmission. When decoding failure is detected, which is generally through cyclic redundancy check (CRC) in a practical system, $\overline{|LLR|}_z$, $z = 1, 2, \cdots, Z$, are calculated for all segments. Only a few segments (with both systematic and corresponding parity bits) that have the lowest $\overline{|LLR|}_z$ are then selected for retransmission. If residual error is detected again, retransmission priorities are given for segments with less retransmission than others and the LLR statistics are calculated for these segments to determine which ones to retransmit. This process continues until no residual error is detected or a hard limit on transmission times is met.

An alternative way to estimate the residual error status of each segment is to use the number of sign changes between LLR value and a priori information. It was found that a decoded bit $x$ has a high error probability when the soft values $L_2^{(i)}(x)$ and $L_{e_1}^{(i)}(x)$ in (2) have different signs [9]. Therefore, we can determine the segments with more number of sign changes of soft values have more errors and select them to retransmit when CRC detects error. Define

$$\begin{aligned} Sgn_z = \; & \tfrac{1}{2}\sum_{j=1}^{l} |\mathrm{sgn}[L_2(X((z-1)l+j))] - \\ & \mathrm{sgn}[L_{e_1}(X((z-1)l+j))]| \end{aligned} \quad (4)$$

as the sum of sign changes between soft values of bits in the $z$-th segment after the final decoding iteration, where sgn[$\cdot$] denotes the operation of obtaining the sign of a real number, $L_{e_1}(X((z-1)l+j))$ is the a priori information of the $j$-th bit in the $z$-th segment for the 2nd component decoder at the final decoding iteration. Then in SSR-based TC-HARQ using sign changes, several segments with the highest $Sgn_z$ are regarded to have more residual errors than others, and are selected to retransmit when error occurs.

As shown in Fig. 5 in Section V, both methods of the mean absolute LLR value and the number of sign change can effectively detect the residual error status of each segment. Therefore, in this research, we will only use mean absolute LLR for the purpose of the segment selection.

### B. Performance of SSR-Based TC-HARQ

The performance of ARQ strategies are evaluated by reliability and throughput [10]. Suppose each coded frame has $n$ bits in which $k$ are information bits, i.e., the coding rate is $R = k/n$. In SSR-based TC-HARQ, the entire frame is sent at the 1st transmission, and only $W$ segments ($W < Z$) will be sent at each retransmission. In type-I TC-HARQ, the entire frame is sent at each transmission and retransmission. Let $N$ be the specified maximum times of transmissions for each frame, which means each frame is obliged to be accepted after the $N$-th transmission even if error still occurs after decoding. Let $R_c^{(i)}$, $R_d^{(i)}$, and $R_u^{(i)}$ denote the events that a decoded frame contains "no errors," "detected errors," and "undetected errors" after the $i$-th transmission, respectively. Then

$$P(R_c^{(i)}) + P(R_d^{(i)}) + P(R_u^{(i)}) = 1, \quad i = 1, 2, \cdots, N. \quad (5)$$

#### B.1 Reliability

Reliability is expressed in terms of the accepted frame error rate, which is the probability that a decoded frame contains either undetected errors within any of the $N$ transmissions or detected errors after the $N$-th transmission, that is,

$$\begin{aligned} P(E) = & P(R_u^{(1)}) + P(R_d^{(1)}R_u^{(2)}) + P(R_d^{(1)}R_d^{(2)}R_u^{(3)}) + \cdots \\ & P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(N-1)}R_u^{(N)}) + P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(N)}). \end{aligned} \quad (6)$$

Note that the $i$-th transmission and decoding processes do not occur unless the decoding of the 1st, 2nd, $\cdots$, and $(i-1)$-th transmissions were all failed with detected errors, i.e.,

$$P(R_d^{(i)}|\overline{R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)}}) = 0. \tag{7}$$

Therefore,

$$\begin{aligned} P(R_d^{(i)}) &= P(R_d^{(i)}|R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)})P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)}) \\ &+ P(R_d^{(i)}|\overline{R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)}})P(\overline{R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)}}) \\ &= P(R_d^{(i)}|R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)})P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)}) \\ &= P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(i)}). \end{aligned} \tag{8}$$

In addition, we clearly have

$$0 \le P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(i-1)}R_u^{(i)}) \le P(R_u^{(i)}). \tag{9}$$

Plugging (8) and (9) into (6) yields

$$P(R_d^{(N)}) \le P(E) \le \sum_{i=1}^{N} P(R_u^{(i)}) + P(R_d^{(N)}). \tag{10}$$

Since an appropriate choice of CRC generally reduces $P(R_u^{(i)})$ below $10^{-5}$ [11], we may assume that $P(R_u^{(i)})$ is negligible. Therefore,

$$P(E) \approx P(R_d^{(N)}). \tag{11}$$

With the same assumption $P(R_u^{(N)}) \approx 0$, $P(R_d^{(N)})$ can be expressed in the function of $P_b^{(N)}$, the residual bit error rate after the $N$-th transmission and the turbo decoding, as

$$P(R_d^{(N)}) \approx 1-(1-P_b^{(N)})^k \tag{12}$$

where $k$ is the number of systematic bits in a frame.

### B.2 Throughput

Throughput is defined as the ratio between the number of all accepted bits at the receiver and the number of all transmitted bits from the transmitter. The average number of transmissions per frame for SSR can be expressed as

$$\begin{aligned} T_{rs} &= 1 + P(R_d^{(1)}) + P(R_d^{(1)}R_d^{(2)}) + \cdots \\ &\quad + P(R_d^{(1)}R_d^{(2)}\cdots R_d^{(N-1)}) \\ &= 1 + P(R_d^{(1)}) + P(R_d^{(2)}) + \cdots + P(R_d^{(N-1)}) \\ &= 1 + \sum_{i=1}^{N-1} P(R_d^{(i)}) \end{aligned} \tag{13}$$

according to (8). This result is slightly different from that with upper and lower bounds in [12] where detection is not assumed to be error-free. From [12], the two bounds are actually the same for high SNR cases and only diverge slightly in low SNR cases where detection errors may exist.

Consequently, the throughput of SSR, represented by $\eta_s$, can be expressed as

$$\eta_s = \frac{k}{n[1+(T_{rs}-1)W/Z]}. \tag{14}$$

In contrast, it is known that the throughput of type-I TC-HARQ $\eta_t$ is

$$\eta_t = \frac{k}{nT_{rt}} \tag{15}$$

where $T_{rt}$ denotes the average number of transmissions per frame for type-I TC-HARQ, which can be obtained similar to that in (13).

Since the numbers of bits to be retransmitted each time are different for SSR-based TC-HARQ and type-I TC-HARQ, to compare these two schemes, we need to set them with the same maximum number of transmitted bits per frame. Let $N_s$ and $N_t$ denote the specified maximum numbers of transmissions per frame in SSR and type-I TC-HARQ, respectively. Then it needs $\eta_{s_{\min}} = \eta_{t_{\min}} = \eta_{\min}$, that is

$$\eta_{\min} = \frac{k}{n[1+(N_s-1)W/Z]} = \frac{k}{nN_t} \tag{16}$$

$$N_t = 1+(N_s-1)W/Z. \tag{17}$$

While the same maximum throughput can be obtained when both strategies have only one transmission per frame, i.e.,

$$\eta_{\max} = \eta_{s_{\max}} = \eta_{t_{\max}} = \frac{k}{n}. \tag{18}$$

The idea of SSR is, by effectively retransmitting the most "wanted" data to obtain $P_b^{(N)}$ (and hence $P(E)$) close to those of type-I TC-HARQ at each transmission, instead of having the relationship between the maximum numbers of transmissions shown in (17), we can actually achieve

$$T_{rs} \le 1+(T_{rt}-1)Z/W \tag{19}$$

within a certain signal to noise ratio (SNR) range. Hence, putting (19) into (14), we have

$$\eta_s \ge \frac{k}{n\{1+[1+(T_{rt}-1)Z/W-1]W/Z\}} = \frac{k}{nT_{rt}} = \eta_t. \tag{20}$$

That is, by retransmitting less bit to correct the errors, SSR can obtain higher throughput than type-I TC-HARQ. It should be noted that this is true only within a specific SNR range where both SSR and type-I TC-HARQ need to transmit more than once but less than the maximum transmission limit per frame on average. SNRs that are too low or too high will lead the throughputs of both schemes to reach the minimum or maximum values, respectively. The effective SNR range is determined by channel type, maximum number of retransmissions allowed, etc.

### B.3 Buffer Size

For data with multiple transmissions, equal gain combining is employed at the receiver for both SSR and type-I TC-HARQ. Let $x_n$ and $x_c^n$ denote the received copy of bit $x$ and the combined result of bit $x$ after the $n$-th transmission, respectively, then the equal gain combining process can be expressed as

$$x_c^n = \frac{x_c^{n-1}(n-1)+x_n}{n}. \tag{21}$$

That is, after each retransmission, the new received data are first combined with the old combined version and then saved into the same buffer unit. Therefore, with equal gain combining SSR and type-I TC-HARQ have the same receiver buffer size.

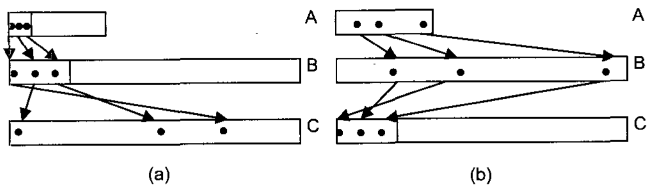


Fig. 2. Modification of SSR for adapting to interleaver: (a) Original SSR, (b) modified SSR.

B.4 Feedback Load

The feedback of SSR-based TC-HARQ is $Z$ bits, the $z$-th ($z = 1, 2, \cdots, Z$) of which denotes whether to retransmit the $z$-th segment, while that of type-I TC-HARQ can be as few as only one bit, denoting whether to retransmit the entire frame or not. Therefore, SSR increases feedback load as compared to type-I TC-HARQ. However, this increase of a few bits is not a large cost. For example, $Z$ is set as 10 in our simulation.

## III. MODIFICATION FOR INTERLEAVER AND FADING CHANNEL

In a practical wireless communication system, an interleaver is always employed before transmission to prevent long blocks of successive errors caused by deep and correlated channel fading. As a result, though one data segment may experience more severe fading than others, with the effect of the interleaver, errors in this segment will be spread out along the entire frame. With the previously discussed SSR strategy, the mean absolute LLR value after decoding does not directly correspond to the residualf error status of a segment at the channel because the order of bits is changed by the interleaver.

Therefore, we further propose a modified SSR strategy to adapt to the use of interleaver. As shown in Fig. 2, 'A,' 'B,' and 'C' denote data before channel encoding, after channel encoding, and after interleaving, respectively. Data in 'C' are to be transmitted over the channel. Previously discussed SSR is shown in Fig. 2(a). Since a segment of data in 'A' and 'B' are scattered in 'C' due to interleaving, they will not experience the correlated fading. Apparently, retransmitting a data segment in 'B' may not get the best possible result. In our modified SSR as shown in Fig. 2(b), instead of retransmitting a segment in 'B,' we retransmit a set of data in B that can form a segment in 'C' so that they experience the correlated fading and might have been severely corrupted together. The positions of this set of data are completely determined by the interleaver and the segment number in 'C.' As a result, this modification in SSR adds no burden for feedback.

Let the order of original information bits be $s_1 = [1, 2, \cdots, n]$, turbo coding rate be $1/r$, then the positions of systematic bits after encoding are $s_2 = [1, r+1, \cdots, (n-1)r+1]$. Let $\beta(i)$ represent the position of bit $i$ after interleaving, then the ordered positions of systematic bits after interleaving are $s_3 = [sort\{\beta(1), \beta(r+1), \cdots, \beta((n-1)r+1)\}]$, where $sort\{\cdot\}$ represents the function of sorting a sequence in increasing order. $s_1$, $s_2$, and $s_3$ denote the positions of information (systematic) bits at 'A', 'B', and 'C' in Fig. 2, respectively. We modify SSR by using $s_3$ as the order of systematic bits to segment data, instead of using $s_1$ in original SSR.

To get the $LLR$ information of bits in $s_3$, let $\beta^{-1}(i)$ represent the position of bit $i$ after deinterleaving, then the positions of bits in $s_3$ after deinterleaving are $s_4 = [\beta^{-1}(s_3[1]), \beta^{-1}(s_3[2]), \cdots, \beta^{-1}(s_3[n])]$. And the positions of bits in $s_4$ after turbo decoding are $s_5 = [(s_4[1] + r - 1)/r, (s_4[2] + r - 1)/r, \cdots, (s_4[n] + r - 1)/r]$. $s_4$ and $s_5$ correspond to the positions of systematic bits at 'B' and 'A' in Fig. 2(b), respectively. Finally, instead of (3), the following equation is used for estimating the residual error status of segments and conducting retransmission.

$$\overline{|LLR|}_z = \frac{1}{l} \sum_{j=1}^{l} |L_2(X(s_5[(z-1)l + j]))|. \qquad (22)$$

We evaluate the effect of this modification by simulations. Each data frame with 1000 bits is uniformly divided into 10 segments. 24-bit CRC with generator polynomial $g_c(x) = (143000003)_{oct}$ [13] is concatenated with a rate-1/3 turbo code with generator polynomial $g_t(x) = (1, 15/13)_{oct}$ [13] for one frame. The intra-frame interleaver defined in [13] is employed before transmission. The popular block fading channel model [14] is considered, where the fading coefficients are assumed to be constant over each segment and vary from one segment to another. The fading envelopes of different segments follow Rayleigh distribution. So that different segments may have different channel gains. The number of transmission for each frame is limited to 3. The entire frame is transmitted at the 1st transmission; 5 segments out of 10 are retransmitted at the 2nd transmission upon request; the rest of the frame with no retransmission yet is retransmitted at the 3rd transmission when error still occurs.

This setting ensures that the BER of original and modified SSR are of zero difference statistically, since the modification only causes difference at the 2nd transmission, if the 2nd transmission fails, the 3rd transmission will be conducted and after 3 transmissions the retransmitted bits of the two strategies are the same. Fig. 4 shows throughput as a function of $E_b/N_0$ for original and modified SSR, where $E_b$ is the average received energy per bit and $N_0$ is the double side noise power spectral density. As expected, modified SSR has distinct improvement over original SSR, because the modified SSR strategy can retransmit more bits severely corrupted in the 1st transmission. Table 1 gives the specific number of transmissions in the experiment, where 1000 frames are considered. It is shown that modified SSR saves the use of the total number of transmissions as compared to original SSR, and thus improves throughput. Apparently that this modified strategy shall result in the highest improvement under block fading channel, some improvement under Rayleigh fading channel, but the same performance under AWGN channel, as compared to original SSR.

## IV. WCDMA SYSTEM OVERVIEW

WCDMA is one important standard of the 3G mobile communication systems [13], [15]. In this paper, the modified SSR is applied to the downlink of the system. The baseband system diagram is shown in Fig. 3.

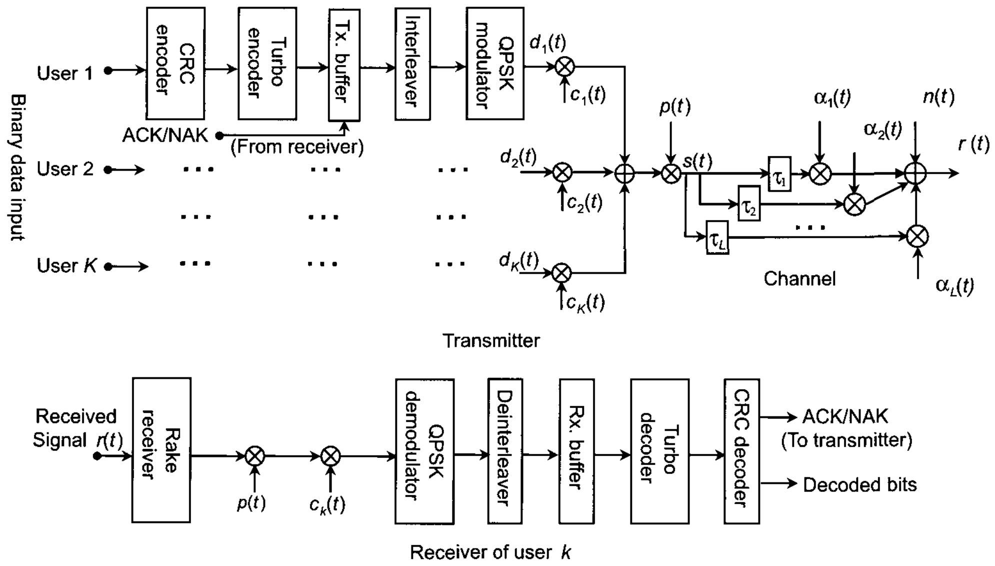


Fig. 3. System diagram of TC-HARQ in WCDMA.

Table 1. Numbers of transmissions for original and modified SSR (1000 frames transmitted, block fading channel).

| $E_b/N_0$ (dB) | −1 | 0 | 1 | 2 | 3 | 4 | 5 | 6 | 7 |
|---|---|---|---|---|---|---|---|---|---|
| No. Tx. modified SSR | 3000 | 2935 | 2683 | 2192 | 1852 | 1465 | 1263 | 1123 | 1019 |
| No. Tx. original SSR | 3000 | 2985 | 2863 | 2442 | 1984 | 1547 | 1287 | 1131 | 1026 |

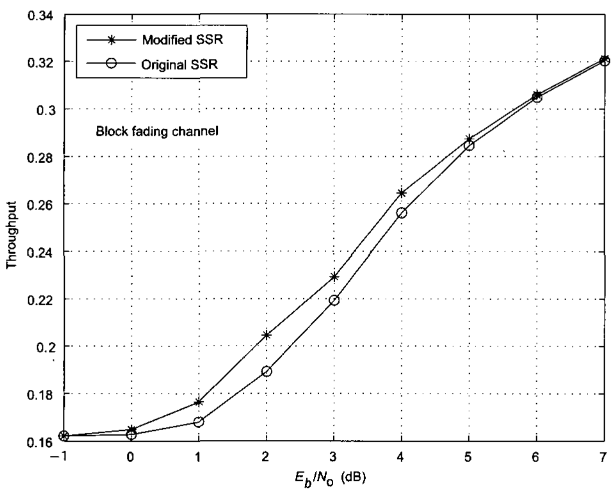


Fig. 4. Throughput comparison of original and modified SSR.

CRC parity bits are attached for error detection, and turbo coding is applied for error correction. The transmission buffer stores the coded data and will either transmit a new frame or retransmit data of the old frame according to acknowledgment (ACK) or negative acknowledgment (NAK) from the receiver. For SSR-based TC-HARQ, the feedback information will be $Z$ bits, indicating which segments to be retransmission. In order to avoid the existence of continuous errors under deep fading, the data after encoding are interleaved so that neighboring bits experience different fading conditions. Quadrature phase shift keying (QPSK) modulation is then employed. The data of different users are spread with specific orthogonal variable spreading factor (OVSF) codes for identifying each other. Each base station uses one identical pseudo-random noise (PN) code sequence $p(t)$ for scrambling the sum of the spread data from different users.

Assume the system has $K$ users, let $d_1(i)$, $d_2(i)$, $\cdots$, and $d_K(i)$ denote the data sequence of user 1, 2, $\cdots$, and $K$ before spreading, respectively. Then the transmitted signal at the base station is [16]

$$s(t) = p(t) \sum_{k=1}^{K} \sqrt{P_k} \times d_k(i) \times c_k(t - i \times SF \times T_{chip}). \quad (23)$$

$P_k$ is the transmission power assigned to user $k$; $c_k(t)$ is the OVSF code sequence used by user $k$; $SF$ is the spreading factor; and $T_{chip}$ is the time duration of one chip. The OVSF codes used by different users compose an orthogonal set $\{c_1(t), c_2(t), \cdots, c_K(t)\}$, which satisfies

$$< c_i(t), c_j(t) >= \begin{cases} 0 & i \neq j \\ SF & i = j \end{cases} \quad (24)$$

where $< \cdot >$ denotes the inner product of two vectors.

The received signal can be represented as

$$r(t) = \sum_{l=1}^{L} e^{j[\omega_c(t-\tau_l)+\phi_l]} \alpha_l(t - \tau_l) s(t - \tau_l) + n(t) \quad (25)$$

where $L$ is the number of paths; $\omega_c$ is the carrier frequency; $n(t)$ is white Gaussian noise with zero mean and variance $\sigma^2 = N_0/2$; $\tau_l$ is the relative delay of the $l$-th path; $\alpha_l(t)$ is the channel gain of the $l$-th path, the envelope of which follows Rayleigh distribution; and $\phi_l$ is the phase shift of the $l$-th path, which is uniformly distributed over $[0, 2\pi)$.

Table 2. Power delay profile of the channel model.

| Rel. delay (nsec) | 0 | 200 | 800 | 1200 | 2300 | 3700 |
|---|---|---|---|---|---|---|
| Avg. power (dB) | 0 | −0.9 | −4.9 | −8.0 | −7.8 | −23.9 |

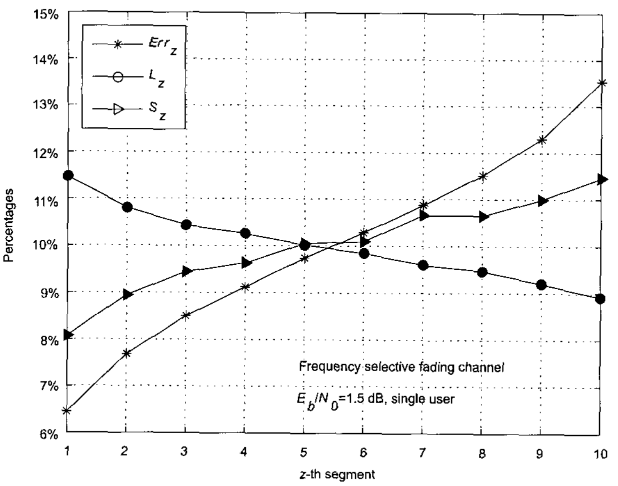


Fig. 5. Residual errors and corresponding mean absolute LLR value and number of sign changes after turbo decoding in WCDMA.

We consider a 6-path frequency selective fading channel model for WCDMA [17]. The power delay profile of the channel is given in Table 2, which shows the relative delay and average power of each path. In the simulation, the delay of each path is rounded to the nearest integer multiples of chip duration. Perfect Rake receiver is assumed to deal with the multipath fading, with 3 fingers correlating to the 3 strongest paths, whose relative delay and average power are assumed available. Maximum ratio combining is employed for combining these paths. The received signal is then descrambled by the original PN code sequence $p(t)$. The desired data of user $k$ is resolved by despreading with $c_k(t)$. The data after demodulation and de-interleaving are passed into the receiving buffer for turbo and CRC decoding. Once CRC detects error, SSR-based TC-HARQ is applied to determine which segments to be retransmitted.

## V. EXPERIMENTAL RESULTS

We evaluate the performance of SSR-based TC-HARQ in WCDMA by simulations. To demonstrate the effectiveness of SSR in determining the residual error status of different segments in WCDMA, we first perform an experiment, where 200 frames are transmitted once at $E_b/N_0 = 1.5$ dB in the single-user case. The $SF$ is set to 32. Frame and segment sizes and codes used are the same as those specified in Section III. Frequency selective fading channel model as defined in Table 2 is applied. Fig. 5 gives the percentages of residual errors, mean absolute LLR value, and number of sign changes of each segment, where segments in each frame are ordered with increasing number of errors. $Err_z$, $L_z$, and $S_z$ are defined as

$$Err_z = \frac{e_z}{\sum_{i=1}^{10} e_z} \times 100\% \qquad (26)$$

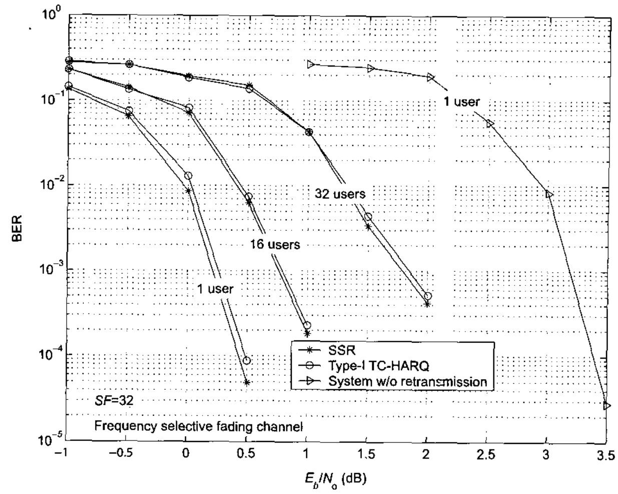


Fig. 6. BER of SSR and type-I TC-HARQ in WCDMA.

$$L_z = \frac{\overline{|LLR|_z}}{\sum_{i=1}^{10} \overline{|LLR|_z}} \times 100\% \qquad (27)$$

$$S_z = \frac{Sgn_z}{\sum_{i=1}^{10} Sgn_z} \times 100\% \qquad (28)$$

where $e_z$, $\overline{|LLR|_z}$, and $Sgn_z$ are the number of errors, the mean absolute LLR value, and the number of sign changes of the $z$-th segment averaged over all frames, respectively. Fig. 5 clearly shows some segments do have more errors than others and SSR-based TC-HARQ could effectively pick them out in WCDMA. A larger mean absolute LLR value or less number of sign changes corresponds to less residual errors in a segment. It is also shown that the mean absolute LLR value and the number of sign changes give similar effects on estimating the residual error status of segments. Therefore, we employ only SSR-based TC-HARQ using mean absolute LLR value in later simulations.

The BER and throughput of SSR and type-I TC-HARQ are then compared in WCDMA. Cases of single, 16, and 32 users are considered. The transmission limit per frame of type-I TC-HARQ is set to 2, while that of SSR is set to 3. The minimum and maximum throughput of 0.162 and 0.324, respectively, can then be achieved by both TC-HARQ strategies according to equations (16) and (18). SSR-based TC-HARQ transmits the entire frame at the 1st transmission, selects 5 segments out of 10 to retransmit at the 2nd transmission upon request, and retransmits the rest 5 segments at the 3rd transmission. Type-I TC-HARQ transmits the entire frame in both transmission and retransmission. The BER performance of SSR and type-I TC-HARQ in WCDMA are shown in Fig. 6. Since SSR effectively selects the most "wanted" data for correcting errors at retransmission, it obtains similar BER as type-I TC-HARQ in all single and multiple user cases. 32-user and 16-user cases exhibit worse performance than single-user case due to the multi-user interference. The BER of single-user system without retransmission is also plotted as a reference.

The throughput per user of SSR and type-I TC-HARQ in WCDMA are given in Fig. 7, which shows the performance in three categories. We take the single-user case as an example. In low $E_b/N_0$ range ($\leq$ 0.5 dB), both strategies have to con-

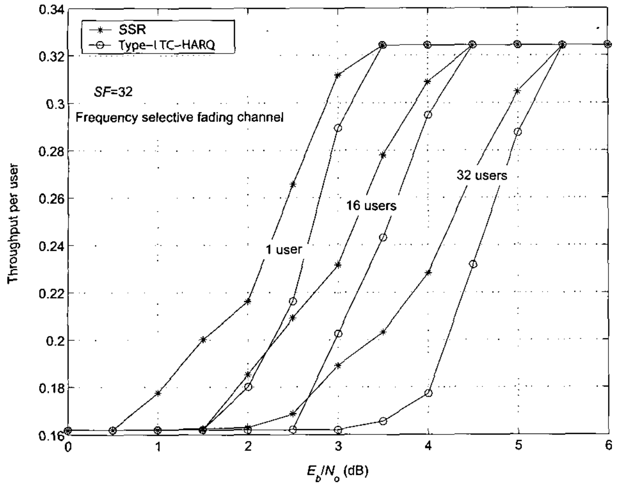


Fig. 7. Throughput per user of SSR and type-I TC-HARQ in WCDMA.

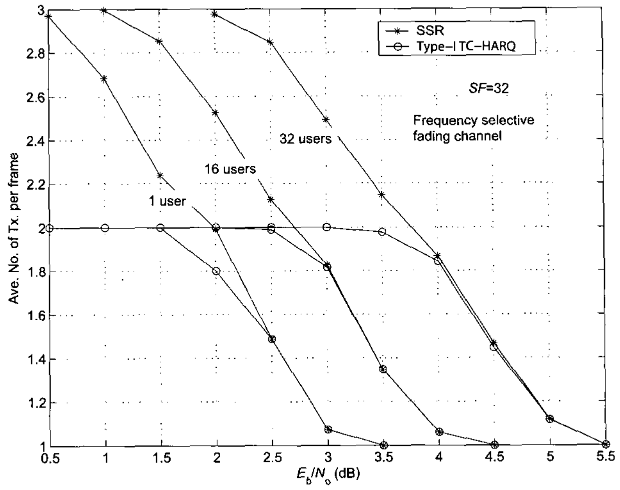


Fig. 8. Average number of transmissions per frame of SSR and type-I TC-HARQ in WCDMA.

sume all possible transmissions and may still have errors. So the same minimum throughputs are obtained by both strategies. In medium $E_b/N_0$ range (0.5–3.5 dB), significant improvement is obtained by SSR, since for many times type-I TC-HARQ needs to retransmit the entire frame to correct errors while by efficiently estimating the positions of segments containing more residual errors, SSR retransmits only the most "wanted" segments of the frame and can correctly decode the data frame. We find that both strategies reach the same maximum throughput in higher $E_b/N_0$ range ($\geq$3.5 dB). This is because one transmission generally is enough to decode all data correctly in such channel condition. This can also be verified from Fig. 6 that the BER of single-user system without retransmission reaches reasonable low value in this range. The throughput in 16 and 32 user's cases demonstrate similar manner as that in single-user case, with the only difference that higher $E_b/N_0$ ranges are required due to the multi-user interference.

It is noticed that the $E_b/N_0$ range for the throughput in Fig. 7 is different from that for the BER in Fig. 6. This is because the throughput will go above the minimum value only when some frames do not require all the number of transmissions, which must have been error free before the final transmission. So the BER should be low enough when the throughput is higher than the minimum value, and thus the $E_b/N_0$ range is higher in Fig. 7 than that in Fig. 6, for corresponding cases.

In the simulation of Figs. 6 and 7, the transmission limit per frame of type-I TC-HARQ is set to 2, while that of SSR is set to 3. So the average number of transmissions per frame of SSR will be larger than that of type-I TC-HARQ. We show in Fig. 8 the actual average number of transmissions per frame for both schemes. It can be seen that SSR does have more transmissions than type-I TC-HARQ, however, the ratio between the number of transmissions of SSR and type-I TC-HARQ is much less than the ratio of 3/2 (obtained at the maximum transmission limit) most of the time, especially in the effective range where SSR achieves much higher throughput than type-I TC-HARQ, such as, 1.5–3.5 dB in single-user case. This is because for most cases SSR can locate and correct the errors at the 2nd transmission, there is no need for the 3rd transmission, though it is available upon request. It is this character of similar number of transmissions but only several segments per retransmission that makes SSR advantageous in effective $E_b/N_0$ ranges.

It should be noted that, the significant improvement in throughput of SSR also requires a few more feedback information than type-I TC-HARQ. With a pre-defined retransmission strategy, type-I TC-HARQ only feeds back one single bit '1' or '0' to indicate 'ACK' or 'NAK.' For SSR with $Z$ segments, $Z$ bits feedback are required, the $z$-th of which being '0' or '1' indicates whether to retransmit the $z$-th segment or not. However, this is a minor cost and is much less in magnitude than the scheme in [4].

The SSR idea can be conveniently generalized to different strategies for specific environments and applications. In the simulation of this paper, we consider each frame to be uniformly divided into 10 segments in SSR, and 5 segments are selected to retransmit each time upon request. However, the number of segments per frame can vary, and the sizes of different segments within a frame can be different, according to any given knowledge on data format, channel condition, and so on. Also, the retransmission strategy is flexible, and the number of segments to retransmit each time can be dynamic based on the estimated residual error status of segments. For example, if after the 1st transmission, it is estimated that all the segments in this frame are very severely corrupted, we can retransmit the entire frame at the 2nd transmission; but if after the 2nd transmission, it is estimated that only a couple of segments have low $\overline{|LLR|}_z$ values, we may retransmit these one or two segments and then correct all errors. It can be noticed that SSR is also able to work directly with type-II TC-HARQ by retransmitting the redundant parity information only for those severely corrupted segments, instead of retransmitting parity bits for all segments.

## VI. CONCLUSIONS

In this paper, the principle of SSR-based TC-HARQ is presented first. Then the performance of SSR is analyzed in terms of reliability and throughput. Later, a modification on SSR for adapting to interleaver under correlated fading channel is proposed. Finally, the modified SSR is applied to the WCDMA system and compared with type-I TC-HARQ under frequency selective fading channel. In all single and multiple user cases, simulation results show that SSR leads to significant improvement in throughput than type-I TC-HARQ with little compromise in BER.

## ACKNOWLEDGMENTS

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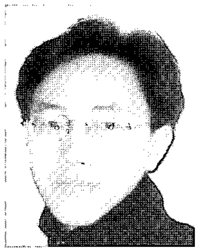

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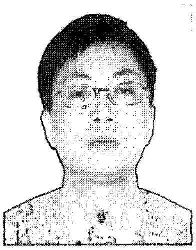

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