Joint Decorrelating Channel and Data Estimation for Space-Time Spreading Systems

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I. Abstract

In this paper, we propose a new joint channel and data detection scheme based on the superimposed training technique for space-time spreading systems. The proposed scheme enhances the performance by eliminating the multiple access interference from both the channel and data estimation by employing two decorrelators: channel and data decorrelators. On a frequency-selective fading channel, our simulation results show that the proposed scheme outperforms other conventional joint channel and data estimation techniques. Moreover, unlike other conventional detection techniques, our proposed estimation technique is shown to achieve the full system diversity.

II. Introduction

Wireless communications for voice and data transmission are currently undergoing very rapid development. Signal processing techniques are incorporated in many of the emerging wireless systems in order to provide advanced services such as multimedia transmission. Code-division multiple-access (CDMA) is seen as one of the generic multiple-access schemes in the second and third generations of wireless communication systems [1]. Despite its promises, CDMA systems have fundamental difficulties when utilized in wideband wireless communications. As the system bandwidth increases, there are more resolvable paths with different delays. Hence, the received CDMA signals suffer from interchip interference (ICI), causing significant cross correlation among users' signature waveforms.

Supporting the expected high data rates required by wireless internet and high-speed multimedia services is one of the basic requirements in broadband mobile wireless systems. In the design of such systems, two critical performance and capacity limiting factors are multipath fading and multiuser interference (MUI) [2]. It has been shown that multiple-input multiple-output (MIMO) based-systems are able to combat fading and improve the system capacity [3],[4]. Within the context of MIMO systems, space-time coding (STC) techniques have been introduced where multiple transmissions from the transmitter side are combined with appropriate signal processing at the receiver to provide diversity gain. Examples of STC techniques are space-time trellis coding (STTC) [5] and space-time block coding (STBC) [6],[7]. It is known that, for the same number of transmit and receive antennas, both STTC and STBC normally achieve the same spatial diversity. However, despite the low complexity they offer, STBCs do not offer any coding gain [8],[9].

In MIMO systems, channel estimation plays a crucial role in determining the system performance. In conventional training-based approaches, a distinct training sequence (known to the receiver) is time-multiplexed with the data sequence and transmitted from the corresponding antenna [1] which will reduce the spectral efficiency. Additionally, for fast-fading channels, such approach is not efficient since training has to be performed repeatedly, which reduces the available time for data transmission [10]. An alternative technique is to estimate the channel while the information signals are being transmitted. This technique is known as blind channel estimation [11]. The major advantage of this technique is the improved bandwidth utilization for fast-fading channels. In semi-blind channel estimation, the channel is estimated not only using the known data in the transmitted signal and its corresponding observation but also using the observation corresponding to the unknown data [12]. More recently, a superimposed training-based approach has been explored where a distinct training sequence is added to the data sequence before modulation and transmission from the corresponding antenna [10]. Thus, this estimation technique has an advantage of not reducing the transmission data rate.

Recently, significant research efforts have aimed at the integration of the superimposed training-based technique with space-time spreading (STS) systems. For example, in [13], the authors employed the training-based technique on STS system with dual transmit and dual receive diversity. The channel estimation was based on employing distinct pilot spreading codes on STS signals transmitted from different antennas. Then, with the help of these pilot signals, the channel coefficients were estimated using a conventional rake receiver. The obtained channel estimates were then used to combine the received signals in a rake-like ST combiner to get the data estimates. This type of joint data and channel estimation suffers from multiple access interference (MAI) and thermal noise effects, especially when employed in loaded environment. In light of this, we propose an enhanced joint channel and data detection scheme based on the superimposed training-based

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approach. The proposed scheme enhances the performance by eliminating the MAI contribution from both the channel and data estimation through employing two decorrelators: channel and data decorrelators.

The remainder of this paper is organized as follows. The following section describes the system model. Channel and data estimations are discussed in sections III and IV respectively. Simulation results are provided in section V. Finally, the conclusion is given in section VI.

III. System Description

The transmit diversity system considered in our work consists of two transmit and one receive antenna. The system block diagram for the k^{th} user is shown in Fig. 1, where real valued data symbols using binary-phase-shift keying (BPSK) baseband modulation and real valued spreading are assumed [14]. We consider the original STS scheme proposed in [15]. As seen in Fig. 1, following the STS, two pilot spreading codes are assigned to each user for the purpose of channel estimation. Each pilot sequence is added (superimposed) to the STS signal transmitted from the corresponding antenna. We also consider uplink asynchronous transmission from K users over frequency-selective slow-fading channel within a data block duration of M symbols. The symbol duration is assumed to be $T_s = UT_b$, where T_b is the bit time-duration and U is the number of bits per symbol which is two in our case [15]. Then the received complex low-pass equivalent signal is given by

$$r(t) = \sum_{k=1}^{K}\sum_{l=0}^{L-1}\sum_{m=0}^{M-1} h_{1l}^{k}\Bigg[\sqrt{\frac{\rho_p}{2}}P_{k1}(t-mT_s-\tau_k-\tau_l)+ \sqrt{\frac{\rho_d}{2}}(b_{k1}[m]c_{k1}(t-mT_s-\tau_k-\tau_l)+b_{k2}[m]c_{k2}(t-mT_s-\tau_k -\tau_l))\Bigg] + h_{2l}^{k}\Bigg[\sqrt{\frac{\rho_p}{2}}P_{k2}(t-mT_s-\tau_k-\tau_l)+\sqrt{\frac{\rho_d}{2}}(b_{k2}[m] \times c_{k1}(t-mT_s-\tau_k-\tau_l)-b_{k1}[m]c_{k2}(t-mT_s-\tau_k-\tau_l))\Bigg] + n(t). \quad (1)$$

where ρ_p and ρ_d represent the pilot and data signal to noise ratios respectively. $b_{k1}[m]$ and $b_{k2}[m]$ are the odd and even k^{th} user data bits within the m^{th} symbol duration. In (1), $c_{k1}(t)$ and $c_{k2}(t)$ are the k^{th} user's data spreading sequences with processing gain UN, where $N = T_b/T_c$ represents the number of chips per bit duration and T_c is the chip-duration. The two pilot spreading sequences, $P_{k1}(t)$ and $P_{k2}(t)$, assigned to the k^{th} user have a period of UT_b. We assume that the pilot and data spreading sequences are mutually orthogonal. τ_k represents the transmit delay of the k^{th} user signal which is assumed to be multiple of chip periods and $\tilde{\tau}_l$ is the multipath delay ($\tilde{\tau}_l = lT_c$). h_{il}^{k}, $i = 1, 2$, is the channel coefficient corresponding to the k^{th} user, l^{th} path from the i^{th} transmit antenna to the base station, and L is the total number of resolvable paths. We also consider a time-invariant channel over the time-duration of a data block, while experiencing independent fading for different data blocks. The noise $n(t)$ is Gaussian with zero mean and unity variance. At the receiver, the received signal is first sent to a channel estimator, where the path gain estimates $\{W_{1l}^{k}, W_{2l}^{k}\}$ are evaluated. Then the STS signals are detected using the estimated path gains. The receiver structure will be explained in more details in the following sections.

IV. Channel Estimation

In the channel estimator, the received signal is chip-shape filtered, sampled at a rate $1/T_c$, and accumulated over an observation interval of $(UN + \tau_{max} + L - 1)$ chips corresponding to the first symbol in the received data block of the K-user system. The $(\tau_{max} + L - 1)$ samples are due to the maximum multipath delay (i.e. the delay of the L^{th} multipath component) corresponding to the user with the maximum transmit delay, τ_{max}. We have chosen to employ the first symbol observation interval in the channel estimation since it is the portion in the received data block with the least MAI and intersymbol interference (ISI) contributions compared with the other received symbols.

Let $\mathbf{y}[0]$ denote the observation vector containing all the samples related to the STS symbols transmitted by the K users within the observation interval, then

$$\mathbf{y}[0] = \mathbf{C}[0]\mathbf{H}\mathbf{b}[0] + \mathbf{C}[1]\mathbf{H}\mathbf{b}[1] + \mathbf{n}[0] \quad (2)$$

where $\mathbf{C}[0] = [\mathbf{C}_1[0], \mathbf{C}_2[0], ..., \mathbf{C}_K[0]]$ represents the code matrix of the K users corresponding to the current received symbols within the observation interval. $\mathbf{C}_k[0]$, $(k = 1, 2, ..., K)$ is a $((UN + L - 1 + \tau_{max}) \times (U + 2)L)$ matrix containing the pilot and data sequences of user k associated with the L multipath components. Similarly, $\mathbf{C}[1] = [\mathbf{C}_1[1], \mathbf{C}_2[1], ..., \mathbf{C}_K[1]]$ represents the code matrix of the K users corresponding to the following received symbols within the observation interval. $\mathbf{C}_k[1]$, $(k = 1, 2, ..., K)$ is a $((UN + L - 1 + \tau_{max}) \times (U + 2)L)$ matrix constructed by the multipath pilot and data sequences of user k associated with the following STS symbol of the k^{th} user within the current observation duration. The second term in the right-hand side of (2) represents the interference due to the following STS symbols of the K users in the system.

Also, in (2), $\mathbf{H}$ represents the channel impulse response of the K users system within the observation interval defined by

$$\mathbf{H} = diag\{\mathbf{H}_1, \mathbf{H}_2, ..., \mathbf{H}_K\}$$

where

$$\mathbf{H}_k = [\mathbf{H}_k^T(0), \mathbf{H}_k^T(1), ..., \mathbf{H}_k^T(L-1)]^T, k = 1, 2, ...K,$$

and the superscript T denotes transpose. $\mathbf{H}_k(l), (l = 0, 1, ..., L - 1)$ is determined according to the employed STS scheme defined in [15] in the case of two transmit antennas which is modified to include the effect of pilot transmissions as follows

$$\mathbf{H}_k(l) = \begin{bmatrix} h_{1l}^k & 0 & 0 & 0 \\ 0 & h_{2l}^k & 0 & 0 \\ 0 & 0 & h_{1l}^k & h_{2l}^k \\ 0 & 0 & -h_{2l}^k & h_{1l}^k \end{bmatrix}.$$

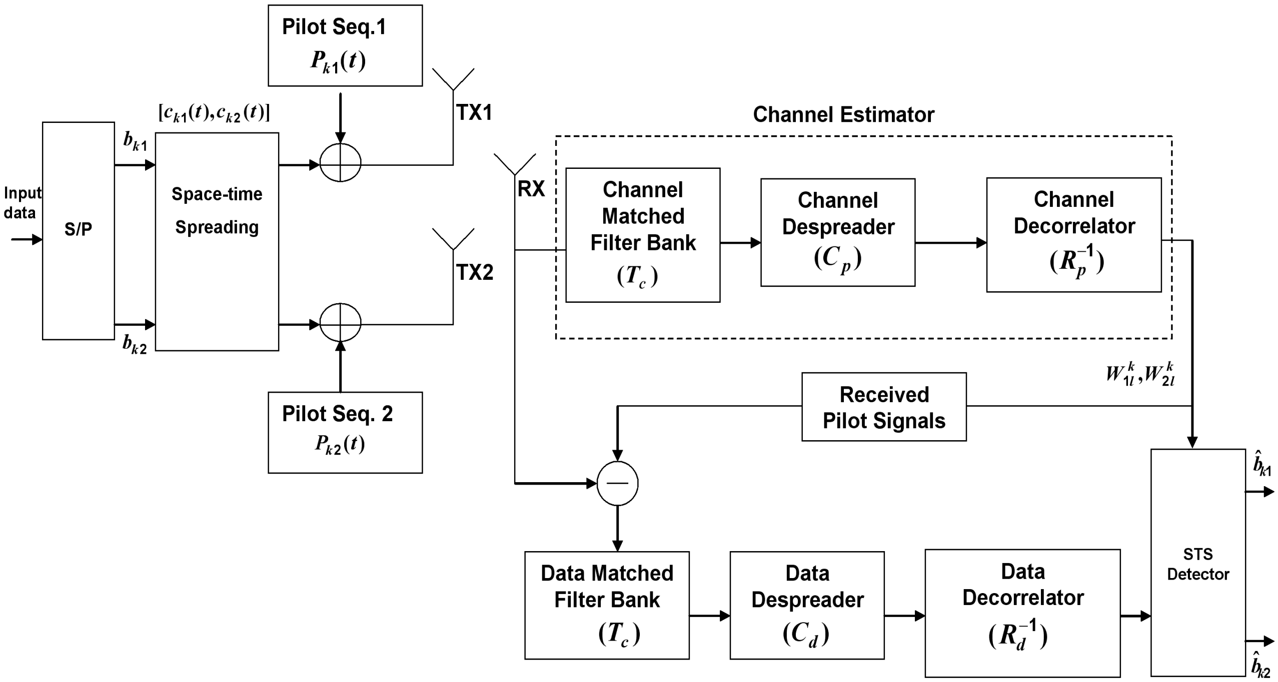

Fig. 1. Block diagram of pilot-sequence-assisted STS transmission system.

The transmitted data vector by the K users during the m^{th} symbol duration, $\mathbf{b}[m]$, is given by

$$\mathbf{b}[m] = [\mathbf{b}_1^T[m], \mathbf{b}_2^T[m], ..., \mathbf{b}_K^T[m]]^T, m = 0, 1, \ldots, M-1$$

where

$$\mathbf{b}_k[m] = \left[\sqrt{\frac{\rho_p}{2}}, \sqrt{\frac{\rho_p}{2}}, \sqrt{\frac{\rho_d}{2}} b_{k1}[m], \sqrt{\frac{\rho_d}{2}} b_{k2}[m]\right]^T , k = 1, 2, ..., K.$$

Finally, in (2), $\mathbf{n}[0]$ is a $((UN+L-1+\tau_{max})\times 1)$ vector that represents the noise samples received within the observation interval which have zero means and unity variances. From (2), the received signal can be represented in a more compact form as

$$\mathbf{y}[0] = \mathbf{C_p}\mathbf{H_p}\mathbf{b} + \mathbf{n}[0] \tag{3}$$

where

$$\mathbf{C_p} = [\mathbf{C}[0], \mathbf{C}[1]],\ \mathbf{H_p} = \mathbf{I}_2 \otimes \mathbf{H},\ \mathbf{b} = [\mathbf{b}[0]^T, \mathbf{b}[1]^T]^T,$$

and $\otimes$ denotes the *Kronecker product* operation [16]. Then the sampled received vector, $\mathbf{y}[0]$, is despread using the pilot-data code matrix $\mathbf{C_p}$ prior to channel estimation, to give

$$\mathbf{y_c}[0] = \mathbf{R_p}\mathbf{H_p}\mathbf{b} + \mathbf{N_{cc}} \tag{4}$$

where $\mathbf{R_p} = \mathbf{C_p}^H\mathbf{C_p}$, is the pilot-data cross correlation matrix and $\mathbf{N_{cc}}$ is modeled as $N_c(\mathbf{0}, \mathbf{R_p})$ (zero mean complex Gaussian vector with covariance $\mathbf{R_p}$) and H denotes Hermitian transpose. Note that, the authors in [13] have based their channel estimation on the channel despreader output, $\mathbf{y_c}[0]$ which implies that the channel estimation in [13] treats the multiuser signal and ISI as a background noise. That is, the error signal in the channel estimates will be affected by the presence of ISI, MAI and thermal noise. Also, to simplify the analysis, the self interference was neglected in [13], relative to the MAI, due to the large number of users. In order to improve the performance, we propose to employ a decorrelator detector for channel estimation after the channel despreader. This will improve the quality of the estimation by removing the effect of MAI from the channel estimates. In this case, the output of the channel decorrelator detector is given by

$$\mathbf{y_d}[0] = \mathbf{H_p}\mathbf{b} + \mathbf{N_{cd}} \tag{5}$$

where $\mathbf{N_{cd}}$ is $N_c(\mathbf{0}, \mathbf{R_p}^{-H})$. The cross correlation matrix inversion is based on the pseudo-inverse or the *Moore-Penrose* generalized inverse [17]. The first $KL(U+2)$ elements are then chosen from the decorrelator output vector, $\mathbf{y_d}[0]$, for estimating the channel coefficients within the assigned data block to give

$$W_{1l}^k = \sqrt{\frac{\rho_p}{2}} h_{1l}^k + e_{1l}^k,$$
$$W_{2l}^k = \sqrt{\frac{\rho_p}{2}} h_{2l}^k + e_{2l}^k, k = 1, 2, ..., K; l = 0, 1, ..., L-1 \tag{6}$$

where e_{1l}^k and e_{2l}^k represent the errors in the channel estimates.

V. Data Detection

Prior to data detection and by using the estimated channel coefficients of the K users, the effect of the received pilot sequences are removed from the received signal defined in (1). Then, similar to the channel estimation procedure, the received signal is filtered, sampled at a rate $1/T_c$, and accumulated over an observation interval of $(UN + \tau_{max} + L - 1)$ chips corresponding to the m^{th} symbol of the received data block. Using vector notation, the chip matched filter output, $\mathbf{g}[m]$ can be expressed as

$$\mathbf{g}[m] = \mathbf{C_d H_d b_d} - \mathbf{P_d E} + \mathbf{n}[m], m = 0, 1, ..., M-1 \quad (7)$$

where $\mathbf{C_d}$, $\mathbf{P_d}$ and $\mathbf{H_d}$ represent the data code, pilot code and channel matrices corresponding to the K-user system within the observation interval. $\mathbf{E}$ is the channel estimation error vector and $\mathbf{b_d}$ is the transmitted data vector. After sampling the received signal, the data matched filter output, $\mathbf{g}[m]$, is correlated with the data code matrix, $\mathbf{C_d}$ as follows

$$\mathbf{g_c}[m] = \mathbf{R_d H_d b_d} - \mathbf{C_d}^H \mathbf{P_d E} + \mathbf{N_{dc}} \quad (8)$$

where $\mathbf{R_d} = \mathbf{C_d}^H \mathbf{C_d}$, represents the data cross correlation matrix. Also, $\mathbf{N_{dc}}$ is modeled as $N_c(\mathbf{0}, \mathbf{R_d})$. It is clear from (8) that the data correlator output, $\mathbf{g_c}[m]$, suffers from MAI. In this case, the output of the data correlator (despreader) is applied to a linear mapper defined by the inverse of the data sequences' cross correlation matrix, $\mathbf{R_d}^{-1}$,

$$\mathbf{g_d}[m] = \mathbf{H_d b_d} - \mathbf{Q_d E} + \mathbf{N_{dd}} \quad (9)$$

where $\mathbf{Q_d} = \mathbf{R_d}^{-1} \mathbf{C_d}^H \mathbf{P_d}$ and $\mathbf{N_{dd}}$ is modeled as $N_c(\mathbf{0}, \mathbf{R_d}^{-H})$. Finally, the first ULK elements of the decorrelator output vector, $\mathbf{g_d}[m]$, are combined with the channel estimates defined in (6) for detecting the data symbols. Considering the first user as the desired user, the decision variables for the first user's odd and even bits are

$$\hat{b}_{11}[m] = \sum_{l=0}^{L-1} Re\{W_{1l}^{1*}(\mathbf{g_d}[m])_{2l+1,1} - W_{2l}^{1*}(\mathbf{g_d}[m])_{2l+2,1}\}, \quad (10)$$

$$\hat{b}_{12}[m] = \sum_{l=0}^{L-1} Re\{W_{2l}^{1*}(\mathbf{g_d}[m])_{2l+1,1} + W_{1l}^{1*}(\mathbf{g_d}[m])_{2l+2,1}\} \quad (11)$$

where $(\mathbf{g_d}[m])_{a,1}$, $(a = 1, 2, ..., UL)$ is the a^{th} element of the decorrelator output vector, $\mathbf{g_d}[m]$.

VI. Simulation Results

In this section, we investigate the bit-error rate (BER) performance of the direct sequence CDMA (DS-CDMA) system using STS and two transmit and one receive antenna over frequency-selective slow-fading channel. We consider the case of imperfect channel estimation at the receiver side. Throughout our simulations, all users are assigned Walsh code sequences of length 128 chips for the pilot and data sequences. We also consider asynchronous transmission over multipath fading channel with $L = 2$ paths per transmit antenna per user. The delay among user signals, (τ_k, $k \in \{1, 2, ..., K\}$), are assumed to be multiple of chip periods within the symbol interval.

Fig. 2 shows the BER performance for different joint channel and data estimation techniques: (i) perfect channel estimation; (ii) joint data and channel estimation without MAI removal [13] (conventional technique); (iii) conventional channel estimation followed by decorrelating data detection (only MAI removal from the data estimates); and (iv) the proposed joint decorrelating data and channel estimation technique. The comparison with the conventional technique is based on the closed form for the probability of error derived in [13]. The results show that the proposed scheme achieves an excellent match with the perfect channel estimation case for $\rho_p = 40$dB. The results also show the effect of MAI removal

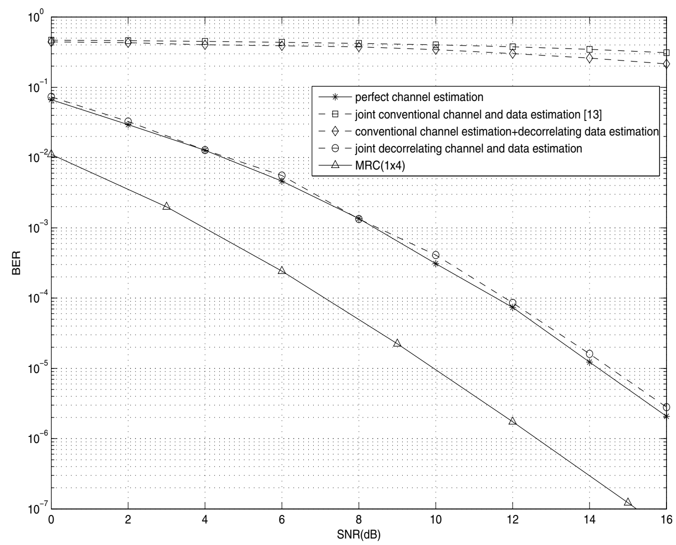

Fig. 2. Comparison between joint data and channel estimation techniques for the 5-user system with L=2 paths and ρ_p=40dB.

from both the channel and data estimates on the system performance. The third system shows a slight improvement over the conventional technique [13], due to the MAI removal from the data estimates but still affected by the imperfect channel estimation. With MAI removal from both the channel and data estimates, the proposed joint decorrelating channel and data estimation technique outperforms the other estimation techniques and achieves the highest convergence to the perfect channel estimation case. For reference, we include the BER performance of the maximal-ratio-combiner (MRC) with four diversity branches. Compared with other conventional estimation techniques, one can notice that the proposed scheme is capable of maintaining the full system diversity at high pilot-to-noise ratio (PNR).

Fig. 3 shows the effect of the PNR on the proposed system performance. Compared with the perfect channel estimation case, we can notice that the accurate estimations are achieved for PNR greater than 30 dB.

VII. Conclusion

We have developed a new channel and data estimation technique for space-time spreading systems. This technique has the advantage of multiple-access interference removal from both the channel and data estimates by employing a decorrelator prior to channel estimation. We have compared our estimation technique with other conventional estimation

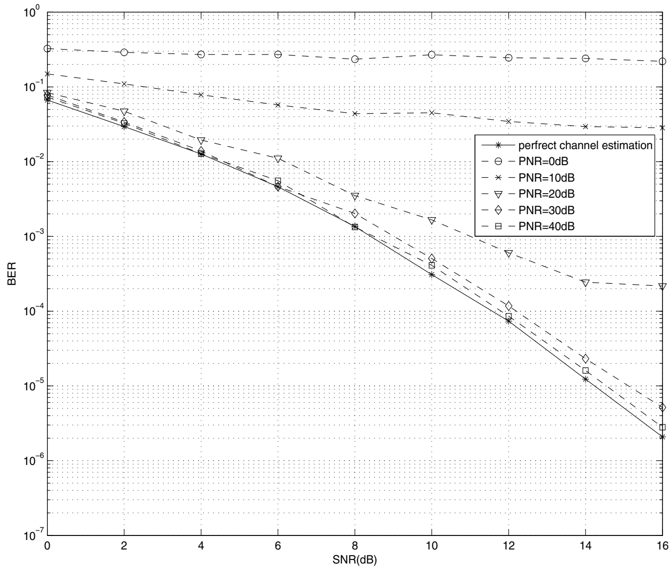

Fig. 3. Effect of PNR on the BER performance of the proposed estimation technique for the 5-user system with L=2 paths.

techniques. Through simulations, we have shown that the proposed channel/data estimation technique is robust to channel estimation errors where it can achieve the highest convergence to the perfect estimation case.

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