the first multipath matched filter output of the lth symbol for the kth user can be expressed as

$$\begin{aligned}\mathbf{y}_{k,1}(l) &= \int_{\tau_{k,1}+(l-1)T}^{\tau_{k,1}+lT} \mathbf{r}(t)s_k(t-\tau_{k,1})dt \\ &= \mathbf{a}_{k,1}e_{k,1}b_k(l) + \sum_{j=2}^{J} \mathbf{a}_{k,j}e_{k,j}b_k(l-\tau_{k,j})\rho^s_{k,k,1,j} \\ &\quad + \sum_{i=1(i\neq k)}^{K} \sum_{j=1}^{J} \mathbf{a}_{i,j}e_{i,j}b_{i,j}(l-\tau_{i,j})\rho^s_{k,i,1,j} + \mathbf{n}_{k,1}(l)\end{aligned} \tag{2}$$

where $\rho^s_{k,i,j,j'} = \int_{\tau_{k,j}}^{\tau_{k,j}+T} s_k(t-\tau_{k,j})s_i(t-\tau_{i,j'})dt$ is the cross-correlation between the spreading waveforms of the jth multipath of the kth user and the jth multipath of the ith user, $\mathbf{n}_{k,1}(l) = \int_{\tau_{k,1}+(l-1)T}^{\tau_{k,1}+lT} \mathbf{n}(t)s_k(t - \tau_{k,1})dt$, and T is the symbol interval. The first multipath beamformer output of the lth symbol for the kth user can be expressed as

$$\begin{aligned}z_{k,1}(l) &= \mathbf{w}^H_{k,1}(l)\cdot\mathbf{y}_{k,1}(l) \\ &= \mathbf{w}^H_{k,1}(l)\mathbf{a}_{k,1}e_{k,1}b_k(l) + \sum_{j=2}^{J} e_{k,j}b_k(l-\tau_{k,j})\rho^s_{k,k,1,j}\rho^a_{k,k,1,j}(l) \\ &\quad + \sum_{i=1(i\neq k)}^{K} \sum_{j=1}^{J} e_{i,j}b_{i,j}(l-\tau_{i,j})\rho^s_{k,i,1,j}\rho^a_{k,i,1,j}(l) + \mathbf{n}_{k,1}(l)\end{aligned} \tag{3}$$

where $\mathbf{w}_{k,1} = [w_{k,1,1}, w_{k,1,2}, ..., w_{k,1,m}]^T$ is the $M \times 1$ weight vector for the first multipath of the kth user, $\rho^a_{k,i,1,j}(l) = \mathbf{a}^H_{i,j} \cdot \mathbf{w}_{k,1}(l)$ is the cross-correlation between the array response vector of the kth user at the first multipath and the weight vector of the ith user at the jth multipath, and $n_{k,1}(l) = \mathbf{w}^H_{k,1}(l)\cdot\mathbf{n}_{k,1}(l)$. The weight vector $\mathbf{w}_{k,1}$ is updated by using least mean square (LMS) algorithm as follows:

$$\mathbf{w}_{k,1}(l+1) = (\mathbf{w}_{k,1}(l)/\|\mathbf{w}_{k,1}(l)\|) + \omega\varepsilon_{k,1}(l)\mathbf{y}_{k,1}(l) \tag{4}$$

where μ is the step size and $\varepsilon_{k,1}(l) = \hat{z}_k(l) - z_{k,1}(l)$ is the error signal. The reference signal $\hat{z}_k$ is the sign of the equal gain combiner output ($\Sigma^J_{j=1} z_{k,j}$) when training sequences are not available, or known sequence when the training sequence is available. In the proposed structure, we use the partial parallel interference cancellation (PIC) based on symbol reliability. Here, we discuss two methods to estimate the proper fractional weight. Using the first method (method (i)), the fractional weight of partial IC based on symbol reliability is estimated as follows:

$$p_{k,j}(l) = \exp\{-(\varepsilon_{k,j}(l)^2)\} \tag{5}$$

The symbol having high reliability provides a smaller mean square error (ε^2) than the symbol having low reliability. Therefore, we give high fractional cancellation weight to the symbol having high reliability. The symbol having lower reliability may degrade the performance of the system owing to incorrect IC. Incorrect choice of symbols adds to interference instead of cancelling it. The partial interference cancelled signal for the kth user can be expressed as

$$\begin{aligned}&\mathbf{r}_k(t) = \\ &\mathbf{r}(t) - \sum_{i=1(i\neq k)}^{K} \sum_{j=1}^{J} \mathbf{w}_{i,j}e_{i,j}s_i(t-\tau_{i,j})p_{i,j}(t-\tau_{i,j})\hat{z}_i(t-\tau_{i,j}) \\ &= \sum_{j=1}^{J} \mathbf{a}_{k,j}e_{k,j}s_k(t-\tau_{k,j}b_k(t-\tau_{k,j}) \\ &\quad + \sum_{i=1(i\neq k)}^{K} \sum_{j=1}^{J} \{e_{i,j}s_i(t-\tau_{i,j})(\mathbf{a}_{i,j}b_{i,j}(t-\tau_{i,j}) \\ &\qquad - \mathbf{w}_{i,j}p_{i,j}(t-\tau_{i,j}))\} + \mathbf{n}(t)\end{aligned} \tag{6}$$

The interference cancelled signal is despreaded, beamformed, the equal gain combined, and decided as follows:

$$\hat{b}_k(l) = \mathrm{sign}\left[\mathrm{Re}\sum_{j=1}^{J}\left\{\mathbf{w}^H_{k,j}\left(\int_{\tau_{k,j}+(l-1)T}^{\tau_{k,j}+lT} \mathbf{r}_k(t)s_k(t-\tau_{k,j})dt\right)\right\}\right] \tag{7}$$

In addition, the fractional cancellation weight can be modified to reduce the degrading effect of the partial interference cancellation. The fractional weights of the highly reliable symbol and the unreliable symbol should be set at 1 and 0, respectively. Therefore, for the second method (method (ii)), we modify the fractional cancellation weight as follows:

$$p'_{k,j}(l) = \begin{cases} 1 & \text{if } p_k(l) \geq TH_h \\ p_{k,j}(l) & \text{if } TH_h > p_k(l) \geq TH_l \\ 0 & \text{if } p_k(l) < TH_l \end{cases} \tag{8}$$

Simulation results and discussion: We carried out simulations to evaluate the performances of the proposed methods over a Rayleigh fading channel. The mean squared value of Rayleigh distribution is 1/8. The number of users is 16 and the random spreading waveform with processing gain $N = 32$ is used. The number of antennas is 4, 6 and 8. The DOA separation between users is 2°. Fig. 2 shows the average BER against received E_b/N_0 for a single path. The upper threshold (TH_h) and lower threshold (TH_l) are 0.82 and 0.42, respectively. The beamformer in Fig. 2 denotes the average BER at the beamformer output. Fig. 2 shows that the IC is required to cancel the residual interferences, which are unsuppressed in the beamformer. In addition, methods (i) and (ii) outperform a combined smart antenna and partial PIC structure based on weight correlation [2] regardless of the number of antennas, as the partial IC based on symbol reliability prevents performance degradation from incorrect choice of symbol. Also, method (ii) outperforms method (i) since method (ii) reduces the degrading effect of the partial IC of method (i). It is noticeable that the proposed combined structure improves performance by preventing, or reducing, the degrading factors of the IC by using properly estimated fractional weights in the initial cancellation stage, and leads to an additional improvement when the number of cancellation stages is increased.

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Location-aware long-life route selection in wireless *ad hoc* networks

Dongkyun Kim, Chai-Keong Toh and Yanghee Choi

A new scheme is proposed for determining a long-life route based on node location information from the GPS (global positioning system) and the radio transmission range of the nodes. The proposed scheme achieves a more stable route than schemes that use the shortest path, avoiding frequent route reconstructions.

Introduction: Although many routing protocols have recently been proposed for wireless *ad hoc* networks, most do not consider route stability [1]. However, the ABR (associativity-based routing) protocol makes routing longevity a criterion for route selection at a receiver [2]. In ABR, to achieve a stable route, each node generates its own periodic beacon signals to maintain associativity information with its neighbours, resulting in a large overhead for the wireless link. In this Letter, we propose a new scheme for determining a long-life route based on node location information obtained from the GPS (global positioning system). The scheme also avoids excessive exchanges of beacon signals among nodes. Although the GPS is used in [3] to acquire a route path, the path selection is still based on the geographically shortest path.

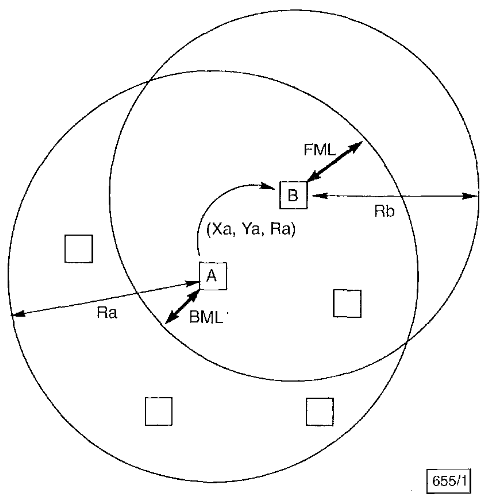

Fig. 1 *FML and BML*

Long-life route selection mechanism: From the GPS, each node obtains its own location information (latitude (x) and longitude (y)), but not that of other nodes. In addition, it is assumed that each node knows its current radio transmission range. Using this information, each node estimates how long the route will last until route failure. A source generates a route discovery packet that contains source position information (X_s, Y_s) and its current transmission range (R_s). This packet is propagated to all neighbouring nodes as in other source-initiated on-demand routing protocols. This is repeated until the destination receives the discovery packet. During the process, node B receiving the route discovery packet from node A performs the following process (Fig. 1).

(i) Using the location information (X_a, Y_a) and radio transmission range (R_a) of the previous node (A) recorded in the route discovery and the location information (X_b, Y_b) of the receiving node (B) obtained from the GPS, calculate the FML (forward movement limit) = R_a – distance(A, B), where distance(A, B) = $\sqrt{[(X_a - X_b)^2 + (Y_a - Y_b)^2]}$.

(ii) Similarly, using the location information and radio transmission range (R_b) of the receiving node (B), calculate the BML (backward movement limit) = R_b – distance(A, B).

(iii) Derive an NML (normalised movement limit) = (FML × BML)/(FML + BML).

(iv) Compare the derived NML with the NML stored in the route discovery packet. Select the minimum value as the propagated NML stored in the route discovery packet.

(v) Forward the route discovery packet to the neighbouring nodes.

An NML should be derived because the FML and BML may be different since time allows all nodes to have different radio transmission ranges due to battery power consumption. When a destination receives a route discovery packet, the packet contains a minimum NML calculated over the route. Since there are many routes between the source and destination, the destination collects multiple route discovery packets traversing different routes. Since a minimum NML contained in the route discovery packet can represent the unstability level of a link over a route, the route with the highest NML among the multiple routes is selected. In addition, since the length of the path is also important for achieving higher throughput as a result of smaller delays, among the routes with the minimum hop counts, the route with the highest NML is selected.

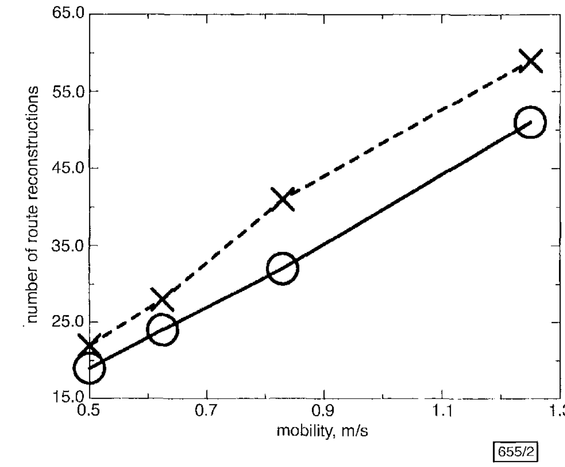

Fig. 2 *Number of route reconstructions against mobility*

○ long-lived route
× pure shortest path route

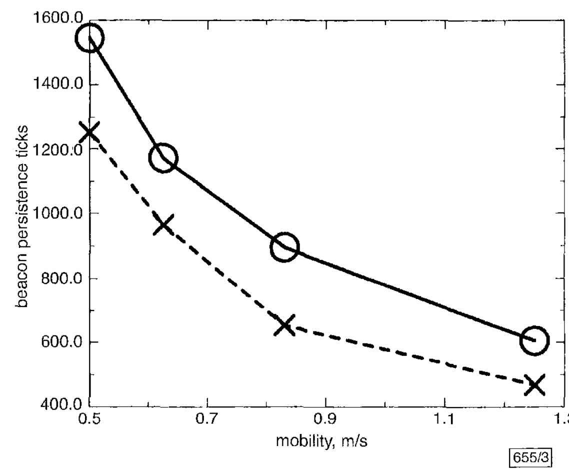

Fig. 3 *Beacon persistence ticks against mobility*

○ long-lived route
× pure shortest path route

Avoiding excessive beacon signals: The proposed scheme does not require the periodic exchange of beacon signals from all adjacent nodes, as is the case for measuring link stability in ABR. In this Letter, since the beacon signals are used to detect route failures on the acquired path, only nodes on the acquired route need generate and exchange their beacon signals to detect a route disconnection, resulting in a lower power consumption than is the case in ABR where all nodes always maintain their link associativity at link level. To detect a route failure, after selecting the best route, the route discovery reply packet is propagated along the reverse path of the selected path and each intermediate node keeps track of its uplink node and downlink node. If there is no beacon signal from the uplink or downlink node during a given period, a node declares a route failure and performs the route discovery process as defined in the ABR protocol [2].

Simulation environment and results: We developed an event-driven simulator suitable for wireless *ad hoc* networks. Initially, 70 nodes were assumed to be spread randomly in the plane (1000 × 1000m). Each node could move randomly in all directions. We assumed that the link propagation delay was 0.01 s and that each node had a radio transmission range of 30m. At first, we compared our approach with the pure shortest path scheme, which does not consider route stability. The number of route reconstructions was measured as the overall node mobility increased. The simulation result in Fig. 2 shows that the frequency of route reconstructions in our scheme is much less than in the pure shortest path scheme.

Assuming a beacon period of one second, a link was considered broken if no beacon signal was received after five seconds. We measured how long a node keeps generating its beacon signals until it experiences a route failure caused by node mobility. It is easy to see that the beacon persistence ticks (the average number of beacon signals between successive route failures) decreases as the node mobility increases. The number of beacon persistence ticks is a measure of route stability. Compared to the pure shortest path scheme, our scheme improves the number of beacon persistence ticks by 20, as shown in Fig. 3.

Conclusion: We have shown that a long-life route selection scheme based on location information and radio transmission range performs much better than the pure shortest route selection scheme in wireless *ad hoc* networks. In addition, thanks to the fact that it is only used for the purpose of detecting route failures, periodic exchange of beacon signals from all ajacent nodes is avoided, resulting in a lower power consumption. Since the GPS may not work in some circumstances, other position estimation schemes need to be considered.

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Transmit diversity system with linear prefilters exhibiting near-optimal performance

Bonghoe Kim and Hwang Soo Lee

A transmit diversity system with liner prefilters exhibiting near-optimal performance is proposed. Without the need for additional antennas, a significant gain can be obtained by increasing the prefilter length and adopting optimal MAP equalisation. As the prefilter length increases, the system performance is improved and closely approaches optimal performance. At a BER of 10^{-3}, the performance of the proposed system attains optimal performance within 0.5dB.

Introduction: In wireless communication systems, fading effects severely degrade system performance. Such fading effects can be effectively mitigated by adopting diversity techniques. Antenna diversity is a simple example of a diversity technique. Using multiple antennas in a receiver, called receive diversity, signals are received from each antenna which undergo independent fading, and diversity gain can be obtained by combining those signals appropriately [1]. However, it is difficult to apply multiple receive antennas in some applications due to space limitation. In these cases, using multiple antennas in a transmitter instead of a receiver, called transmit diversity, can provide a diversity gain. Several transmit diversity techniques with bandwidth efficiency, where there exists no feedback path, have been proposed. The optimal prefilters for a transmit antenna are found by minimising the variance of the normalised squared Euclidean distance in [2]. A transmit diversity system with linear prefilters and a linear equaliser is developed in [3]. However, the results in [2, 3] focus on the case where the number of antennas is equal to the length of the prefilter. In this Letter, we propose a transmit diversity system with linear prefilters that can achieve near-optimal performance without the need for additional antennas. This performance can be obtained by increasing the length of the linear prefilters and adopting MAP equalisation.

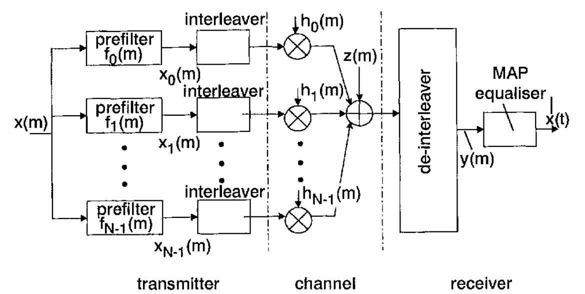

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Fig. 1 *Proposed transmit diversity system with linear prefilters*

Proposed system description: A block diagram of the proposed transmit diversity system with linear prefilters is shown in Fig. 1. We consider N transmit antennas and one receive antenna. We assume that the linear prefilters have a finite impulse response (FIR) and that the following interleavers all have the same pattern. The output symbol of the linear prefilter in the nth antenna branch at time m is

$$x_n(m) = \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} f_n(k)x(m-k) \qquad n = 0, 1, ..., N-1 \tag{1}$$

where K is the length of the prefilter, $f_n(k)$ is the kth coefficient of the nth prefilter, and $x(m)$ is the information data. For each antenna, $x_n(m)$ is interleaved and transmitted over a frequency-nonselective Rayleigh fading channel. The role of the interleaver is to diminish the time correlation between the fading samples and improve the MAP equalisation performance.

The received and deinterleaved symbol is

$$\begin{aligned} y(m) &= \sum_{n=0}^{N-1} h'_n(m)x_n(m) + z'(m) \\ &= \sum_{n=0}^{N-1} \frac{h'_n(m)}{\sqrt{K}} \sum_{k=0}^{K-1} f_n(k)x(m-k) + z'(m) \\ &= \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} \sum_{n=0}^{N-1} h'_n(m)f_n(k)x(m-k) + z'(m) \\ &= \sum_{k=0}^{K-1} g(m,k)x(m-k) + z'(m) \end{aligned} \tag{2}$$

where $h'_n(m)$ is the deinterleaved version of $h_n(m)$ which is complex Gaussian and characterises the nth transmission path from the nth transmit antenna to the receive antenna, $z'(m)$ is the deinterleaved version of $z(m)$ which is complex AWGN with zero mean and two sided noise density $N_0/2$, and $g(m, k) = (1/\sqrt{K}) \Sigma_{n=0}^{N-1} h'_n(m)f_n(k)$ is the equivalent channel representation for the overall system, which includes the linear prefilters, interleavers, transmission channel and deinterleaver. The equivalent channel coefficient $g(m, k)$ is Gaussian because linear combination of Gaussian random variables is also Gaussian. The equivalent channel model is shown in Fig. 2. As shown in Fig. 2, a frequency-nonselective fading channel for each transmission path is transformed into a frequency-selective fading channel [3]. Since the equivalent channel has memories, an equaliser is required in the receiver side. We consider the MAP equaliser as an optimal equaliser because a MAP equaliser can provide greater gain than an MLSE equaliser when combined with channel coding. However, we do not consider channel coding