1.43%. Using linear interpolation for fading estimation, the BER performances with L = 70, 40 and 10 are shown in the same Figure for comparison. It can be seen that, with L = 70 and 40, irreducible error floors occur at BER = 5×10^{-2} and 5×10^{-3}, respectively, whereas the proposed technique can reduce the BER to 2×10^{-4}, achieving substantially better performance improvements. With L = 10, linear interpolation seems to perform better than the proposed technique, but the required bandwidth redundancy is now 10%, instead of 1.43%.

With a faster normalised fading rate of $f_D T = 10^{-2}$, the BER performances of the signal with L = 35, 20 and 5 are shown in Fig. 2. It can be seen that the performances are slightly worse than in the slow fading environment of Fig. 1. Here again, no error floor occurs under the conditions tested. The results obtained using linear interpolation are also shown for comparison. It can be seen that, with linear interpolation and L = 35 and 20, the error floors occur at BER = 5×10^{-2} and 5×10^{-3}, respectively. If the proposed technique is used, the BERs are substantially lower. Both the results in Figs. 1 and 2 indicate that the proposed technique is a bandwidth-efficient method. In a bandwidth limited system, the proposed PSA technique is obviously better for fading estimation.

Conclusions: A bandwidth-efficient technique for use in PSA systems has been described and studied. The technique employs both data and pilot symbols to reduce the square of the estimation error. Computer simulation results have shown that, in frequency non-selective Rayleigh fading channels corrupted with AWGN, the technique requires a minimal bandwidth redundancy to transmit the pilot symbols. In a slow fading environment with $f_D T = 5 \times 10^{-3}$, the bandwidth redundancy required to transmit the pilot symbols can be reduced to as low as 1.43%. Moreover, the proposed technique achieves substantially lower error floors.

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Blind RAKE receiver with joint diversity combining and code tracking for DS/SS communication

Jia-Chin Lin

A RAKE receiver achieving joint blind multipath diversity combining and code tracking is proposed. An improved known modulus adaptive algorithm is exploited to perform multipath diversity combining and to support the modified code tracking in the blind mode. Computer simulation results have indicated very attractive behaviour of the proposed technique.

Introduction: Frequency-selective fading can often lead to severe performance degradation in wideband communication systems. Substantial efforts have been made in adaptive equalisation and diversity combining (RAKE) techniques in order to improve receiver performance, but little work has been devoted to optimising code synchronisation systems, although the conventional code tracking loop (DLL) is vulnerable to multipath fading effects [1]. In fact, only when both multipath diversity combining and code tracking are optimised simultaneously for frequency-selective fading effects can overall receiver performance be improved. In this Letter, a technique achieving joint blind multipath diversity combining and a code tracking loop is proposed. Based on this technique, a multipath diversity combiner with the improved known modulas adaptive (KMA) algorithm operates with the modified PN code timing recovery to achieve simultaneous improvement of overall receiver performance with very low computation load. In the modified PN code timing recovery, the timing error signals are independently extracted and then effectively combined in the same fashion as that exploited in the multipath diversity combiner to achieve improved code tracking performance as well.

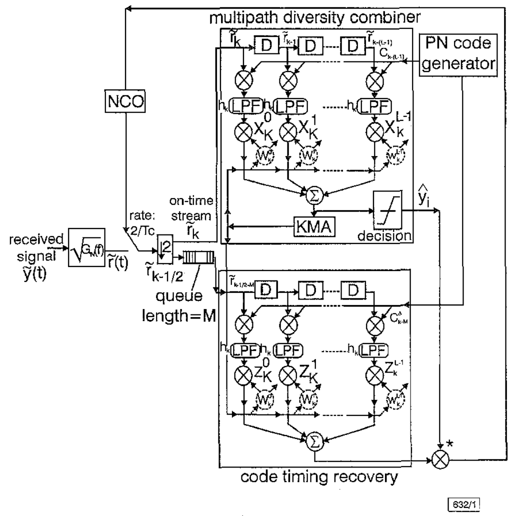

Fig. 1 *RAKE receiver based on proposed technique*

Proposed RAKE receiver: A complete block diagram of the RAKE receiver based on the proposed technique is shown in Fig. 1. The complex representation of the baseband signal at the output of the chip matched filter is

$$\tilde{r}(t) = \sum_{n=0}^{L-1} a_n(t) s(t - nT_c) + n(t) \qquad (1)$$

where $s(t) = \Sigma_{i=0}^{\infty} d_i \Sigma_{l=0}^{M-1} c_l g(t - lT_c - iT_b)$ is the data-modulated PN sequence with raised cosine chip shaping, d_i is the ith information-bearing symbol, c_l is the lth chip value of the PN sequence, T_b and T_c are the symbol interval and the chip duration, respectively, $M = T_b/T_c$ is the processing gain, $g(t)$ is the overall chip shape. The signal $\tilde{r}(t)$ is sampled at twice the chip rate, i.e. sampled at the instants $t_k = (k + \varepsilon_k)T_c$ and $t_{k-\frac{1}{2}} = (k + \varepsilon_k - \frac{1}{2})T_c$, where ε_k is the kth normalised chip timing error, to produce the two parallel sequences: integer-instant or on-time samples $\tilde{r}_k = \tilde{r}(t_k)$, and half-integer-instant samples $\tilde{r}_{k-\frac{1}{2}} = \tilde{r}(t_{k-\frac{1}{2}})$.

Multipath diversity combining technique based on improved KMA algorithm: The integer-instant sample stream

$$\tilde{r}_k = \sum_{n=0}^{L-1} a_n(kT_c) s(kT_c - nT_c + \varepsilon_k T_c) + n_k \qquad (2)$$

is fed into the multipath diversity combiner based on the improved KMA algorithm. Assume first that the code acquisition process has been achieved. On each arm of the multipath diversity combiner, the input samples $\tilde{r}_{k-m}$, $m = 0, 1, ..., L-1$, are cross-correlated with the local PN sequence $c_{k-(L-1)}$, which has been code-acquired, and then passed through the arm filter, h_k, which is a lowpass filter with bandwidth B_b comparable with the symbol rate, $1/T_b$, in order to reject effectively the PN self-noise, multiple access interference, and noise outside the bandwidth of the information-bearing symbols. The input samples x_k^m, $m = 0, 1, ..., L-1$, of the improved KMA algorithm proposed here are taken from the output of the mth arm filter h_k and given as

$$x_k^m = \tilde{r}_{k-m} \times c_{k-(L-1)} * h_k \qquad (3)$$

where $\times$ and $*$ denote the correlation and convolution operators, respectively.

To achieve multipath diversity combining, a multipath diversity combiner with coefficient vector $\mathbf{W}_k = [w_k^0, w_k^1, ..., w_k^{L-1}]^T$ is employed. The multipath diversity combiner output is

$$y_k = \mathbf{W}_k^T \mathbf{X}_k \tag{4}$$

where $\mathbf{X}_k = [x_k^0, x_k^1, ..., x_k^{L-1}]^T$ is the input vector of the improved KMA algorithm and x_k^m, $m = 0, 1, ..., L-1$ are taken from the outputs of the arm filters as shown in eqn. 3. Based on the derivation and discussion presented previously [3], the update procedure can be rewritten as

$$\mathbf{W}_{k+1} - \mathbf{W}_k = -\frac{\mu}{\|\mathbf{X}_k\|^2}\mathbf{X}_k^*[\hat{e}_k^R + j\hat{e}_k^I] \tag{5}$$

and the error signal $[\hat{e}_k^R + j\hat{e}_k^I]$, can thus be approximated as

$$\begin{aligned} \hat{e}_k^R &= \begin{cases} \frac{\tilde{e}_k^R}{(y_k^R)^2} & y_k^R \notin Z_k^R \\ y_k^R - \hat{y}_k^R & y_k^R \in Z_k^R \end{cases} \\ \hat{e}_k^I &= \begin{cases} \frac{\tilde{e}_k^I}{(y_k^I)^2} & y_k^I \notin Z_k^I \\ y_k^I - \hat{y}_k^I & y_k^I \in Z_k^I \end{cases} \end{aligned} \tag{6}$$

where

$$\begin{cases} \tilde{e}_k^R = y_k^R((y_k^R)^2 - R_k^R) & R_k^R = \frac{E\{[d_k^R * h_k]^4\}}{E\{[d_k^R * h_k]^2\}} \\ \tilde{e}_k^I = y_k^I((y_k^I)^2 - R_k^I) & R_k^I = \frac{E\{[d_k^I * h_k]^4\}}{E\{[d_k^I * h_k]^2\}} \end{cases} \tag{7}$$

Z_k^R and Z_k^I are the variable confidence zones of the real and imaginary parts, respectively, and $\hat{y}_k^r$ and $\hat{y}_k^I$ are the real and imaginary parts, respectively, of the decision result of the multipath diversity combiner output.

PN code tracking loop: Here, we describe the operations in the modified PN code tracking loop. The early-late structure is replaced by the very simple digital correlator with the half-integer-instant stream and the code difference stream as its inputs. The incommg delayed (one symbol interval, $T_b = MT_c$, delayed) half-integer-instant samples $\tilde{r}_{k-\frac{1}{2}-M-m}$, $m = 0, 1, ..., L-1$, are cross-correlated with the code difference stream, $c_{k-M}^{\Delta} = c_{k-(L-1)-M} - c_{k-L-M}$. After passing through the arm filter, h_k, the arm error signal z_k^m on the mth arm is generated in the following form: $z_k^m = \tilde{r}_{k-\frac{1}{2}-M-m} \times c_{k-M}^{\Delta} * h_k$. The arm error signal z_k^m on the mth arm has the same error characteristics as those of the conventional delay locked loop (DLL) operated on a single-path channel with AWGN, but here it is corrupted by the corresponding channel tap weight and the information-bearing symbol. The compound error signal, e_k, can be obtained by combining the arm error signals, z_k^m $\forall m$, with the tap weights, w_k^m $\forall m$, which are the same as those exploited in the multipath diversity combiner; then, the data modulation effect on z_k^m compensated for by the decision-directed method:

$$e_k = \mathrm{Re}\left\{\hat{y}_{[k]M}^* \sum_{m=0}^{L-1} w_k^m z_k^m\right\} \tag{8}$$

The effects of channel multipath fading, the carrier phase error and data modulation on the arm timing error signal, z_k^m, can be simultaneously overcome by exploiting multipath diversity combining.

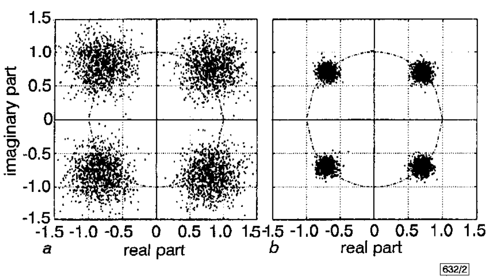

Fig. 2 *Signal constellations after multipath diversity combiner with EKF-based estimator* [2] *and proposed technique*

a EKF when SNR = –5dB
b MKMA when SNR = –5dB

Simulation results: Computer simulation results illustrating the performance of the proposed technique are presented in this Section. Fig. 2 shows the signal constellations obtained by a multipath diversity combiner with the EKF-based estimator [2] and the proposed technique, in the case of a frequency offset $\Delta f/(1/T_c) = 10^{-4}$ and SNR = –5dB. The multipath diversity combiner, aided by the EKF-based estimator [2], and with the proposed technique, can track the carrier frequency offset and cluster the output signal constellation at the right position. It is also obvious that the multipath diversity combiner with the proposed technique can cluster the output signals at the right position much better than can one aided by the EKF-based estimator [2].

Conclusion: A technique of joint blind multipath diversity combining and code tracking is proposed in this Letter. It has been shown that this technique can accomplish multipath diversity combining in the blind mode. The simulation results show that such a technique can certainly achieve simultaneous improvement of multipath diversity combining and code tracking on frequency-selective fading channels.

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Comment

Bit error rate performance of π/4 DQPSK for Nakagami-lognormal channels

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A convergent infinite series has been presented for the bit error performance of π/4 DQPSK for Nakagami-lognormal channels if the fading figure *m* is an integer. The authors show that this infinite series can be replaced by a finite series. A convergent series for the BER is also derived when the fading figure, *m*, is real.

Introduction: In a recent Letter [1], Tjhung and Chai have derived an infinite series for the bit error rate (BER) of π/4 DQPSK for Nakagami-lognormal (NLN) channels provided the Nakagami fading figure is constrained to be an integer. This modulation scheme, differential quadrature phase shift keying, has been adopted for several practical mobile systems. Therefore, evaluation of its performance under general fading and shadowing conditions is useful. The purpose of this Letter is to suggest some improvements to the infinite-series solution.

Theory: The conditional BER, $P_b(e|s)$, derived in [1] utilises a BER expression from [2] which involves the generalised Marcum's Q function and an infinite series of modified Bessel functions. Consequently, eqn. 8 in [1] is an infinite series. However, it possible to replace this by a finite series. From [3], the conditional BER can be written as