CMA/FRACTIONAL-DELAY SEQUENTIAL BEAMFORMING FOR WIRELESS MULTIPATH COMMUNICATIONS

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Abstract— **This paper presents a new adaptive antenna-array structure using CMA/Fractional-Delay sequential blind beamforming for multipath correlated signals in environments, such as indoor confined or underground areas where the multipath problem is more severe than the co-channel interference. The new receiver uses a new synchronization approach for multipath propagation, based on combining a CMA adaptive antenna array and an adaptive fractional time delay estimation filter. This approach is designed to sequentially recover multipath rays based on adaptive time delay estimation for fractional time path arrival delays. The bit-error rate is evaluated for different operating conditions. Simulation results show the promising performance of the new adaptive antenna-array structure.**

Keywords- Adaptive antenna array, beamforming, CMA, LMS, fractional delay, adaptive signal canceller, multipath fading channel, adaptive signal processing.

I. INTRODUCTION

The main cause of degradation of communication quality in harsh confined environments such as underground mines is multipath fading as it is typically more severe than co-channel interference (CCI) [1-2]. To mitigate multipath fading, adaptive antenna arrays (AAA) have been studied [3-4]. Recently many different types of antenna beamforming have been developed. For instance, the constant modulus algorithm (CMA), applied in a blind AAA, is considered a promising method in mobile communications for mitigating multipath fading by effectively suppressing delayed paths and CCI signals [5]. However, suppressing the multipath rays of the desired signal wastes part of the available power and requires additional degrees of freedom. In addition, since these arrival paths are delayed replicas from an identical source, it is desirable to separate and combine the delayed paths instead of suppressing them for received power maximization.

Several methods have been proposed to separate and combine the delayed paths. One of these methods is the subband filter [6] that is based on increasing the signal correlation between the direct path and the delayed paths in each subband. Other methods such as the spatial-domain path-diversity (SPD) systems [7-10], using multibeam AAA for the TDMA system, are proposed. In these methods multiple beams are formed to separate the direct path and the delayed paths by using training sequences.

These aforementioned approaches have certain advantages and limitations, some of them require training sequences and all are trying to solve the problem of multipath propagation, when the time path arrival (TPA) is an integer multiple of the sampling interval. For non-integer TPA, other approaches are used such as a Rake receiver to treat the path arrivals at the chip interval [11] or straightforwardly, over-sampling [12] is employed. Nevertheless, over-sampling will not complicate only the structure of the analog-to-digital converter, but also may not synchronize correctly with the required time path arrival τ.

Generally speaking, when the TPA is a non-integer multiple of the sampling interval, power available in these paths is wasted and fractional time delay estimation (FTDE) is required to overrule the over-sampling solution. FTDE is employed in several signal processing applications which include also time delay difference (TDD) between signals received at two spatially separated antenna arrays (SSAA) [13]. It consists in using a linear interpolator to implement a Fractional Delay Filter (FDF) [13]. An FDF is a type of digital filter designed for band-limited interpolation which can be implemented using an FIR filter based on the Lagrange sinc-interpolation method [14].

Recently we proposed a blind spatial-domain path-diversity beamforming to recover the signal and its integer multiple replicas using jointly CMA and LMS. In this paper, an extension of our recent work is presented when a new synchronization approach for SPD is proposed, based on combining a CMA-AAA [3] and an adaptive FTDE filter based on an FIR sinc-interpolation filter [13] named FD-CMA. It addresses a new approach to Time Delay Estimation (TDE) when the TPA is a fractional value. This approach is designed to sequentially recover multipath rays by using multiple beamformings for received power maximization. The paper is organized as follows. In Section 2, the proposed algorithm is described and mathematical formulations are derived. Section 3 presents the performance of the FD-CMA by computer simulation results, and Section 4 concludes this paper.

II. PROPOSED ALGORITHM

As mentioned above, the proposed algorithm is dedicated to confined and underground environments, where the multipath problem is more severe than CCI.

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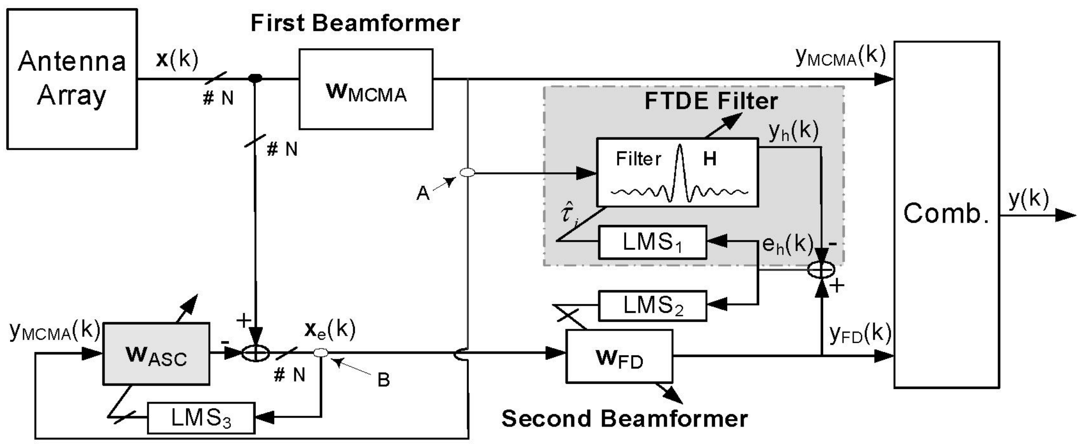

Figure 1. Proposed FD-CMA Algorithm.

- **Signal Model**

To easily clarify the principle of the proposed method, this study is restricted to a two-path channel model. Consider an N-element AAA. The received signal vector is given by:

$$\mathbf{x}(k) = \sum_{i=1}^{2} \alpha_i . s(k - \tau_i) . \mathbf{a}_s + \boldsymbol{\eta}(k). \tag{1}$$

where, $s(k)$ is the desired signal, drawn from alphabet members $A=\{a_1,\ldots,a_m\}$, α_i represents the complex amplitude of the ith path, $\mathbf{a}_s$ represents the propagation vector, $\boldsymbol{\eta}(k)$ is the additive white Gaussian noise vector, $\tau_1 = 0$ (we synchronize with the first path) and $\tau_2 = \tau$ ($0 < \tau < T_s$). For simplicity, the first path is assumed the strongest.

Fig. 1 represent the new structure for spatial-domain path-diversity which resolves multipath signals where TPA is a non-integer multiple of the sampling interval. This method is implemented using two sequential beamformers. The first beamformer is used to estimate the strongest path while its weights ($\boldsymbol{w}_{MCMA}$) are adapted using a Modified-CMA (MCMA) [3]. The second beamformer ($\boldsymbol{w}_{FD}$) is used to estimate the 2nd path coming with a fractional delay and adapted using the least mean square algorithm (LMS_2 in Fig.1). The first beamformer output is fed into an FTDE filter to generate the reference signal used to adapt the weights of the second beamformer. The weight adaptation for both FTDE and $\boldsymbol{w}_{FD}$ filters is nested in the sense that the output of the $\boldsymbol{w}_{FD}$ filter is used as reference signal to the FTDE filter and vice-versa. However, to ensure that the second beamformer detects the 2nd path and not the first one, an adaptive signal canceller (ASC) [15] is used to extract the contribution of the direct path.

The procedure of the proposed algorithm is as follows:

1. The first beamforming weights for the strongest path are estimated using MCMA, and the output of this beamformer is named y_{MCMA}.
2. The contribution of the first estimated path (y_{MCMA}) will be extracted from the received signal by using an ASC; the result of this subtraction is denoted as $\boldsymbol{x}_e$.
3. The filter H is used to delay the estimated path y_{MCMA} by a fractional delay value (τ).
4. The output of this filter H (y_h) is used as a reference signal to estimate the coefficients of the second beamformer for the remaining path using LMS ($\boldsymbol{w}_{FD}$).
5. The second beamforming weights for the 2nd path are estimated using LMS, and the output of this beamformer is named y_{FD}.
6. The output of the second beamformer (y_{FD}) is used as a reference signal to adapt the delay τ by using LMS.

Finally, these two beamformers' output signals $y_{MCMA}(k)$ and $y_{FD}(k)$ are combined to maximize the received power by diversity combining.

- **Modified CMA- AAA**

Given a beamformer weight vector $\boldsymbol{w}_{CMA}(k)$, the output of the beamformer is:

$$y_{CMA}(k) = \mathbf{w}^H_{CMA}(k) \cdot \mathbf{x}(k), \tag{2}$$

The CMA-AAA aims to eliminate the amplitude fluctuations of the array output signal due to the incidence of interferences. Therefore, the cost function to be minimized is represented as:

$$J(\mathbf{w}_{CMA}) = E\left[\left(\left|y_{CMA}(k)\right|^2 - R_{CMA}\right)^2\right], \tag{3}$$

where, $E[\cdot]$ denotes the ensemble mean and R_{CMA} is a constant which depends on the input symbols a. This constant is defined by [3]:

$$R_{CMA} = \frac{E\left[|a|^4\right]}{E\left[|a|^2\right]}, \quad \text{for } a \in A . \tag{4}$$

Since CMA is phase blind, the array output will have an arbitrary phase rotation at the convergence. To address this problem a Modified CMA (M-CMA) is used [3]. In this algorithm, the cost function is divided into real and imaginary parts, as follows:

$$J(\mathbf{w}_{MCMA}) = E\left[\left(\left|y_{MCMA\_R}(k)\right|^2 - R_R\right)^2\right] + E\left[\left(\left|y_{MCMA\_I}(k)\right|^2 - R_I\right)^2\right], \tag{5}$$

where, $$R_R = \frac{E\left[a_R^4(n)\right]}{E\left[a_R^2(n)\right]}, R_I = \frac{E\left[a_I^4(n)\right]}{E\left[a_I^2(n)\right]}, \quad (6)$$

$$a(k) = a_R(k) + j.a_I(k), \quad (7)$$

$$y_{MCMA}(k) = y_{MCMA\_R}(k) + j.y_{MCMA\_I}(k). \quad (8)$$

A stochastic gradient search method can be used to minimize the MCMA cost function by adaptively adjusting the weight vector according to:

$$\mathbf{w}_{MCMA}(k+1) = \mathbf{w}_{MCMA}(k) - \mu \cdot e^*(k) \cdot \mathbf{x}(k), \quad (9)$$

where, the error function is given by:

$$e(k) = e_R(k) + j.e_I(k), \quad (10)$$

$$e_R(k) = y_{MCMA\_R}(k) \cdot (y^2_{MCMA\_R}(k) - R_R), \quad (11)$$

$$e_I(k) = y_{MCMA\_I}(k) \cdot (y^2_{MCMA\_I}(k) - R_I). \quad (12)$$

- **Adaptive Signal Canceller (ASC)**

Once the first path is estimated by M-CMA, its contribution is removed from the received signal vector $\boldsymbol{x(k)}$ using an ASC. By adopting this canceller filter, it becomes possible for the second beamformer to estimate the 2nd path and for the FTDE filter to estimate the delay value of this path. This ASC employs $\boldsymbol{x(k)}$ as a reference signal. It consists of a weight vector updated by LMS where the error vector $\boldsymbol{x_e(k)}$ is given by [15]:

$$\mathbf{x}_e(k) = \mathbf{x}(k) - \mathbf{w}^H_{ASC}(k) \cdot y_{MCMA}(k). \quad (13)$$

The weight vector is given by:

$$\mathbf{w}_{ASC}(k+1) = \mathbf{w}_{ASC}(k) + \mu_3 \cdot \mathbf{x}^H_e(k) \cdot y_{MCMA}(k), \quad (14)$$

where μ_3 is a small positive step-size.

The signal at point A in Fig. 1 is expressed by:

$$y_{MCMA}(k) = \hat{s}(k), \quad (15)$$

By summing the vector $\boldsymbol{x_e(k)}$ at point B in Fig. 1, the result $x_{e\_sum}(k)$ can be expressed by:

$$x_{e\_sum}(k) = \beta(k) \cdot \hat{s}(k-\tau) + \gamma(k). \quad (16)$$

where $\beta(k)$ is a multiplicative factor and $\gamma(k)$ is an additive white Gaussian noise vector. So these two points, A and B, can be regarded as two SSAA. By analogy with the TDE method proposed in [13], the delay between the signals $s(k)$ and $s(k-\tau)$ can be estimated. However, instead of summing directly the signal $\boldsymbol{x_e(k)}$ at point B to construct the reference signal of the FTDE filter, the filter $\boldsymbol{w_{FD}}$ is added to construct the second beamformer to estimate the 2nd path as depicted in Fig.1.

- **Fractional Time Delay Estimation Filter**

Once the first path is estimated by MCMA, it is delayed by an estimated value $\hat{\tau}$ using the fractional delay filter $\boldsymbol{H}$. This filtering is carried out by using the following equations [13]:

$$y_h(k) = y_{MCMA}(k-\hat{\tau}) = \sum_{n=-\infty}^{\infty} sinc(n-\hat{\tau}) \cdot y_{MCMA}(k-n),$$

$$y_h(k) = \sum_{n=-P}^{P} sinc(n-\hat{\tau}) \cdot y_{MCMA}(k-n). \quad (17)$$

where the infinity sign in the summation is replaced by an integer P, that is chosen large enough to minimize truncation error, and where $\hat{\tau}$ is the instantaneous estimated time delay. This delayed signal $y_h(k)$ is the output of an FIR filter $\boldsymbol{H}$ whose coefficients are $sinc(n-\hat{\tau})$ and input is $y_{MCMA}(k)$. A lookup table of the *sinc* function is constructed. It consists of a matrix $\boldsymbol{H}$ of dimension $K \times (2.P+1)$ with element:

$$h_{ij} = sinc(\frac{i-1}{K} - j) \qquad 1 \le i \le K, \quad -P \le j \le P. \quad (18)$$

where K represents the inverse resolution over T_s of the estimated delay $\hat{\tau}$. In our simulation K is taken equal to 200.

The elements of the ith row of matrix $\boldsymbol{H}$ are therefore identical to the samples of a *sinc* function with delays equal to:

$$\tau_i = (i-1)/K. \quad (19)$$

Thus, for a given τ_i (17) can be expressed as:

$$y_h(k) = \mathbf{h}^H_i \cdot \mathbf{u}(k), \quad (20)$$

where $\boldsymbol{h_i}$ is the ith row of the matrix $\boldsymbol{H}$ and $\boldsymbol{u(k)}$ is given by:

$$\mathbf{u}(k) = \left[y_{MCMA}(k) \;\; \ldots \;\; y_{MCMA}(k-(2P+1))\right]^T. \quad (21)$$

The estimated fractional time delay is obtained by gradient descent of the instantaneous squared error $|e_h|^2$ surface in order to locate the global minimum, i.e., using LMS [13]. This method is extended in this work to be applied to complex signals. The estimated gradient is equal to the derivative of $|e_h|^2$ with respect to $\hat{\tau}$. The FTDE algorithm may be summarized as follows. The complex error signal $e_h(k)$ is given by:

$$e_h(k) = y_{FD}(k) - y_h(k), \quad (22)$$

$$e_h(k) = y_{FD}(k) - \sum_{n=-P}^{P} sinc(n-\hat{\tau}(k)) \cdot y_{MCMA}(k-n), \quad (23)$$

where, $$y_{FD}(k) = \mathbf{w}^H_{FD}(k) \cdot \mathbf{x}_{e\_D}(k), \quad (24)$$

$$\mathbf{x}_{e\_D}(k) = \mathbf{x}_e(k-(P+1)). \quad (25)$$

$\boldsymbol{x_e(k)}$ is delayed by $(P+1).T_s$ to be aligned with the output of the filter $\boldsymbol{H}$ that has latency depending on its order value $M=2P+1$.

The estimated time delay can be adapted by minimizing the cost function given by:

$$J(\tau, \mathbf{w}_{FD}) = E\left[e_h(k)\right]^2 = E\left[y_{FD}(k) - y_h(k)\right]^2, \quad (26)$$

The constrained LMS algorithm becomes:

$$\hat{\tau}(k+1) = \hat{\tau}(k) - \mu_1 \cdot \nabla J(\tau, \mathbf{w}_{FD}). \quad (27)$$

where μ_1 is a small step-size.

By differentiating the instantaneous error surface, $e_h^2(k)$, with respect to the estimated time delay we obtain:

$$\frac{\partial(e_h^2(k))}{\partial\hat{\tau}(k)} = \sum_{n=-P}^{P} f(n-\hat{\tau}(k))\cdot y_{MCMA}(k-n)\cdot e_h^*(k) + \sum_{n=-P}^{P} f(n-\hat{\tau}(k))\cdot y_{MCMA}^*(k-n)\cdot e_h(k), \quad (28)$$

where,
$$f(\upsilon)=\frac{\cos(\pi\upsilon)-sinc(\upsilon)}{\upsilon}. \quad (29)$$

Finally, the estimated time delay $\hat{\tau}$ is given by:

$$\hat{\tau}(k+1)=\hat{\tau}(k)-\mu_1\cdot\left[\sum_{n=-P}^{P} f(n-\hat{\tau}(k))\cdot y_{MCMA}(k-n)\cdot e_h^*(k) + \sum_{n=-P}^{P} f(n-\hat{\tau})\cdot y_{MCMA}^*(k-n)\cdot e_h(k)\right]. \quad (30)$$

In our simulations, lookup tables of *cos* and *sinc* functions are constructed for different values of υ and used to calculate $f(n-\hat{\tau}(k))$. At each iteration, the integer part of $(\hat{\tau}_i(k)\cdot K+1)$ is used to locate the *i*th row of the matrix $\boldsymbol{H}$, i.e. $\boldsymbol{h}_i$ that is used to delay the signal $y_{MCMA}(k)$ using (20).

Minimization of the cost function given in (26) is also used to adapt the weight vector of the second beamformer as follows:

$$\mathbf{w}_{FD}(k+1)=\mathbf{w}_{FD}(k)+\mu_2\cdot e_h^*(k)\cdot\mathbf{x}_{e\_D}(k). \quad (31)$$

where μ_2 is a small step-size.

Finally, the signals $y_{MCMA}(k)$ and $y_{FD}(k)$ are added to maximize the power at the receiver output.

III. COMPUTER SIMULATION RESULTS

In this section, simulation results are presented to assess the performance of our proposed FD-CMA beamforming and to compare it with the MCMA beamforming. A two-element array with half-wavelength spacing is considered. A BPSK desired signal is propagated along 4 multipaths to the AA while the interference and noise are simulated as white Gaussian noises. The first path is direct with time path arrival delay $\tau_1=0$. The second and third paths arrive, respectively, with delays τ_2 and τ_3 lower than the sampling interval. And the last path arrives with delay $\tau_4=T_s$. In our simulation environment, the strongest path (τ_1) is extracted using MCMA while the fourth ($\tau_4=T_s$) is estimated using our method presented in [3]. The new proposed method FD-CMA is used to extract the remaining non-integer paths (τ_2 and τ_3). Differential encoding is employed to overcome the phase ambiguity in the signal estimation. The performance studies are carried out with several channel models. The type-A channel is Rayleigh fading with a Doppler shift f_{d1}=20Hz. The type-B channel is Rayleigh fading with a higher Doppler shift f_{d2}=35Hz. The use of these two Doppler frequencies reflects the typical range of the vehicle speed in our underground environment. The BER performance for different Doppler frequencies (f_{d1} and f_{d2}) was studied. The figure of merit is the required SNR to achieve a BER below 0.001.

Figures 2 and 3 illustrate the measured BER performance versus SNR for FD-CMA and M-CMA for channels Type-A and -B for $\tau_2=0.2T_s$ and $\tau_3=0.8T_s$, respectively, at 2.4 GHz. These results show that FD-CMA provides good enhancement and outperforms M-CMA by approximately 4 dB and 5 dB at a required BER=0.001 for both channel types, respectively.

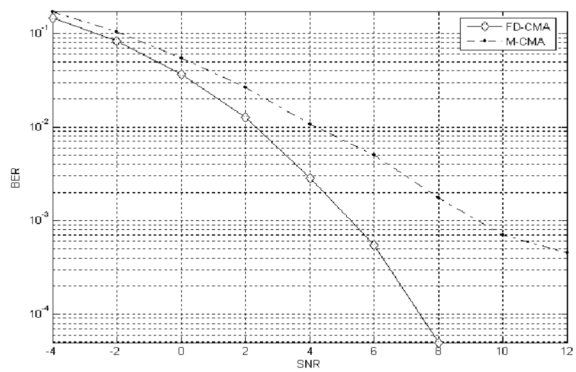

Figure 2. BER performance versus SNR in Type –A Channels for τ_2=0.2 T_s and τ_3=0.8 T_s.

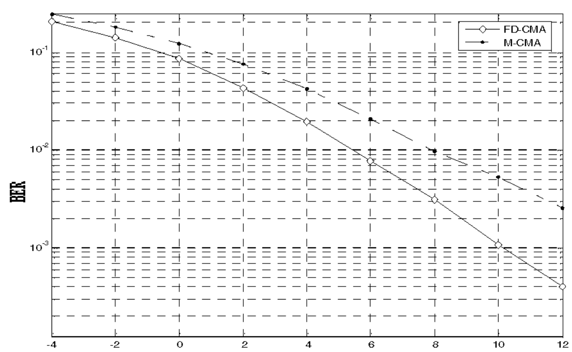

Figure 3. BER performance versus SNR in Type –B Channels for τ_2=0.2 T_s and τ_3=0.8 T_s.

Figures 4 and 5 show the BER performance of the FD-CMA for both channel environments (A and B), respectively, at 2.4 GHz and for different antenna numbers, i.e. N=2 and N=4. The results indicate that the improvement in the BER performance is proportional to the number of elements exploited in the adaptation. From these results, it can be noted that this improvement decreases with the Doppler frequency. Furthermore, with FD-CMA using only two antenna elements, we can obtain almost the same performance with M-CMA using four antenna elements.

IV. Conclusion

In this paper, a new FD-CMA, applied in AAA, has been proposed using sequential blind beamforming for multipath correlated signals in environments, such as underground areas, where the multipath problem is more severe than the CCI. The proposed AAA uses a new synchronization approach for multipath propagation; based on combining a CMA-AAA and adaptive fractional time delay estimation filter. This approach is designed to sequentially recover multipath rays to maximize the received power by looking for all dominant multipaths. In this method, instead of delaying the CMA output by one symbol to look for the ISI, as proposed in [3], this output is delayed by an estimated fractional delay value to recover the arrival paths with delays less than the symbol period. The new approach for TDE is based on the truncated *sinc* fractional delay filter algorithm.

Simulation results show that the proposed FD-CMA receiver outperforms the MCMA in all scenarios and especially at high SNR where the receiver is expected to operate in an underground wireless environment [1].

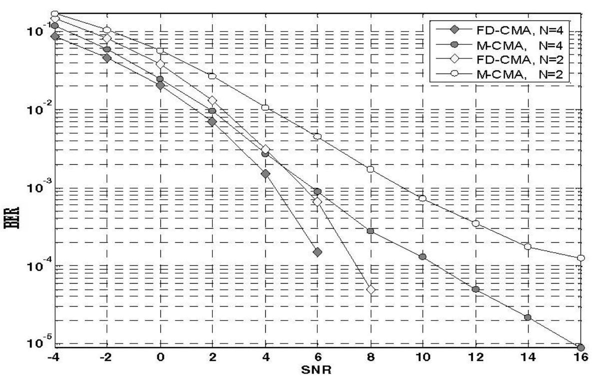

Figure 4. BER performance versus SNR in Type –A Channels when the number of elements is varied.

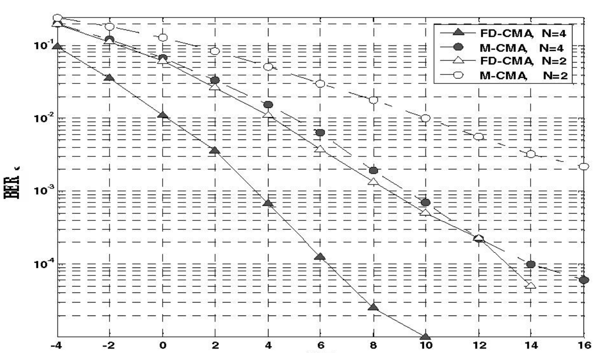

Figure 5. BER performance versus SNR in Type –B Channels when the number of elements is varied.

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