Frequency-domain Adaptive Antenna Array for Single-carrier Uplink Transmission Using Frequency-Domain Equalization

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***Abstract*—The bit error rate (BER) performance of single-carrier (SC) block transmission in a frequency-selective fading channel can be significantly improved by the use of frequency-domain equalization (FDE). However, in multi-cell environment, the BER performance is degraded due to co-channel interference. Adaptive antenna array (AAA) is effective to suppress the co-channel interference. In this paper, we propose frequency-domain adaptive antenna array (AAA) for SC uplink transmission using MMSE-FDE. Since the frequency-domain AAA tries to form an antenna beam pattern based on the minimization of the average interference power, the same array weight can be used for all the frequency components. After the array combining, FDE is performed. In the paper, the BER performance is evaluated by computer simulation to show that the array weight converges in one block even if the normalized LMS algorithm is used.**

Keywords-sigle-carrier transmission, adaptive antenna array, frequency-domain equalization

I. INTRODUCTION *(HEADING 1)*

In the next generation mobile wireless communication systems, high speed and high quality data services are demanded. Since the channel is composed of many propagation paths with different time delays, the bit error rate (BER) performance of single-carrier (SC) transmission significantly degrades due to serve inter-symbol-interference (ISI) [1]. Recently, it was shown that frequency-domain-equalization (FDE) based on minimum mean square error (MMSE) criterion can obtain the frequency diversity gain [2, 3] and hence, improve the BER performance. In the SC transmission with FDE, a symbol sequence is divided into blocks of N_c symbols and then, the last N_g symbols of each block are copied as a cyclic prefix and are inserted into the guard interval (GI) at the beginning of each block, similar to OFDM. In a cellular system, the same carrier frequency is reused in different cells. Co-channel interference from the other cells' users reaches the BS and this degrades the BER performance. If the co-channel interference is suppressed, the distance between the co-channel cells can be reduced, thereby improving the frequency efficiency. One of effective techniques to suppress the co-channel interference is adaptive antenna array (AAA) [4, 5].

In this paper, frequency-domain AAA is proposed for SC uplink (mobile-to-base) transmission with FDE. The proposed frequency-domain AAA tries to form an antenna beam pattern based on the minimization of the average interference power, therefore the same array weight can be used for all the frequency components. A pilot block is transmitted every N–1 data blocks. Using a pilot block, the array weight is updated for frequency-domain array combining for the reception of N–1 data blocks following the pilot block. The array weight can be updated the same number of times as the number of subcarriers using the normalized least mean square (NLMS) algorithm [6] (here, we use the terminology, "subcarrier" for explanation purpose only). After the array combining, MMSE-FDE is applied.

The rest of this paper is organized as follows. Sect. II presents the SC transmission system model using the proposed frequency-domain AAA. In Sect. III, the array weight updating, using NLMS algorithm is described. In Sect IV, the array weights convergence rate is investigated and the BER performance is evaluated by computer simulation to show that the array weight can converge in a block even if the NLMS algorithm is used. Sect. V gives some conclusions.

II. TRANSMISSION SYSTEM MODEL OF FREQUENCY-DOMAIN AAA

A. Transmit signal

Fig.1 shows the transmitter/receiver structure of frequency-domain AAA for single-carrier with FDE. At the transmitter, binary data sequence is transformed into a data-modulated symbol sequence, and divided into blocks of N_c data symbols. For channel estimation, a pilot block is inserted every N–1 data blocks as shown in Fig.2. The last N_g symbols in each block of N_c symbols are copied and inserted as a cyclic prefix into the guard interval (GI) which is placed at the beginning of each block [2, 3].

Symbol-spaced discrete time representation is used throughout the paper. The uth user's transmit symbol sequence $\{\tilde{s}_u(t); t=0\sim N_c-1\}$ is expressed, using the equivalent baseband representation, as

$$\tilde{s}_u(t)=\sqrt{\frac{2E_{s,u}}{T_s}}s_u(t)\ ,(1)$$

where $E_{s,u}$ represents the uth user's transmit signal energy per symbol and T_s is the symbol length.

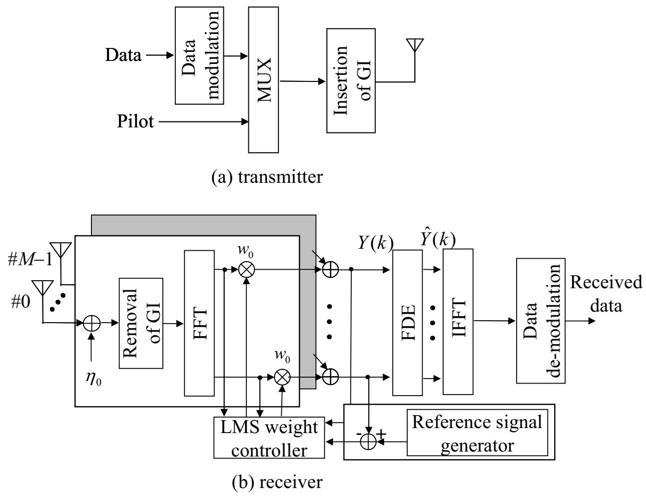


Figure 1. SC transmission system model with joint freqeucny-domain AAA and FDE.

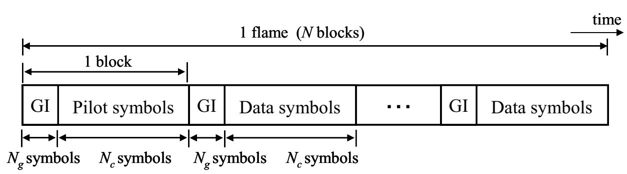


Figure 2. Frame structure.

B. Receive signal

We consider U–1 interfering users from other co-channel cells. The u=0th user is the desired user and the other U–1 users are interfering users. The chip sequence received on the mth receive antenna is expressed as

$$r_m(t)=\sqrt{\frac{2E_{s,0}}{T_s}}\sum_{l=0}^{L-1}h_{0,m,l}s_0(t-\tau_{0,l})+\sum_{u=1}^{U-1}\left\{\sqrt{\frac{2E_{s,u}}{T_s}}i_{u,m}(t)\right\}+\eta_m(t)\ ,(2)$$

where $i_{u,m}(t)$ is the uth interfering user's signal from the other co-channel cell and given by

$$i_{u,m}(t)=\sum_{l=0}^{L-1}h_{u,m,l}s_u(t-\tau_{u,l})\ .(3)$$

$h_{u,m,l}$ is the complex-valued path gain between the uth user and the mth (m=0~M–1) receive antenna, having $\sum_{l=0}^{L-1}E\left[\left|h_{u,m,l}\right|^2\right]=1$ and $\tau_{u,l}$ represents the lth path time delay (l=0~L–1) including the transmit timing offset. $\eta_m(t)$ is a zero-mean complex Gaussian process with a variance of $2N_0/T_s$ with N_0 being the single-sided power spectrum density of the additive white Gaussian noise (AWGN) process.

After the removal of the GI, $r_m(t)$, m=0~M–1, is decomposed into N_c subcarrier components $\{R_m(k);\ k=0\sim N_c-1\}$ by applying N_c-point FFT. The kth subcarrier component $R_m(k)$ associated with the mth (m=0~M–1) antenna is expressed as

$$\begin{aligned}R_m(k)&=\sum_{t=0}^{N_c-1}r_m(t)\exp\left(-j2\pi k\frac{t}{N_c}\right)\\&=H_{0,m}(k)S_0(k)+\sum_{u=1}^{U-1}I_{u,m}(k)+\Pi_m(k)\end{aligned}\ ,(4)$$

where $H_{0,m}(k)$, $S_0(k)$, $I_{u,m}(k)$ and $\Pi_m(k)$ are the kth channel gain, the kth subcarrier component of the transmitted symbol sequence, that of the interference from the uth co-channel cell and that of the noise due to the AWGN, respectively. They are respectively given by

$$\begin{cases}S_0(k)=\sum_{t=0}^{N_c-1}s_0(t)\exp\left(-j2\pi k\frac{t}{N_c}\right)\\H_{0,m}(k)=\sqrt{\frac{2E_{s,0}}{T_s}}\sum_{l=0}^{L-1}h_{0,m,l}\exp\left(-j2\pi k\frac{\tau_{0,l}}{N_c}\right)\\I_{u,m}(k)=\sqrt{\frac{2E_{s,u}}{T_s}}\sum_{t=0}^{N_c-1}i_{u,m}(t)\exp\left(-j2\pi k\frac{t}{N_c}\right)\\\Pi_m(k)=\sum_{t=0}^{N_c-1}\eta_m(t)\exp\left(-j2\pi k\frac{t}{N_c}\right)\end{cases}\ ,(5)$$

C. Frequency-domain array combining

The array weight updating is described in Sect. 3. Let $\mathbf{R}(k)=\left[R_0(k)\ R_1(k)\ \cdots\ R_{M-1}(k)\right]^T$ and $\mathbf{W}=\left[w_0\ w_1\ \cdots\ w_{M-1}\right]^T$ be the received signal vector and the array weight vector, respectively. The array output signal $Y(k)$ is expressed as

$$Y(k)=\mathbf{R}^T(k)\mathbf{W}\ ,(6)$$

where $(\bullet)^T$ represents the transposition. The same array weight is used for all the subcarriers. The reason for this is that the statistical properties of the interference are the same for all the subcarriers.

D. MMSE-FDE

After the array combining, one-tap FDE is performed on $Y(k)$, as

$$\hat{Y}(k)=Y(k)w_{\mathrm{FDE}}(k)\ ,(7)$$

where $w_{\text{FDE}}(k)$ is the MMSE-FDE weight, given by

$$w_{\text{FDE}}(k) = \frac{\left\{\mathbf{H}_0^T(k)\mathbf{W}_q\right\}^*}{\left|\mathbf{H}_0^T(k)\mathbf{W}_q\right|^2 + 2\sigma^2} \quad ,(8)$$

where $\mathbf{H}_u(k) = \left[H_{u,0}(k)\ H_{u,1}(k) \cdots H_{u,M-1}(k)\right]^T$ is the channel gain vector and $2\sigma^2$ is the variance of the noise plus the residual co-channel interference after array combining. N_c-point inverse FFT (IFFT) is applied to $\{\hat{Y}(k); k = 0 \sim N_c - 1\}$ to obtain the time-domain chip sequence $\{y(t); t = 0 \sim N_c - 1\}$ for data demodulation:

$$y(t) = \frac{1}{N_c}\sum_{k=0}^{N_c-1}\hat{Y}(k)\exp\left(j2\pi t\frac{k}{N_c}\right) \quad .(9)$$

III. ARRAY WEIGHT UPDATING

Array weight is updated using the received pilot block. The kth subcarrier component $R_m(k)$ of the received pilot can be expressed as

$$R_m(k) = H_{0,m}(k)P_0(k) + \sum_{u=1}^{U-1} I_{u,m}(k) + \Pi_m(k) \quad ,(10)$$

where $P_0(k)$ is the kth subcarrier component of the transmitted pilot. The antenna beam pattern is formed to minimize the average interference power after the array combining. As described in Sect. II C, the same array weight is used for all the subcarriers. Hence, a fast adaptive algorithm is not necessary since it is possible to update the weights as many as N_c times in the frequency-domain in one block. In this paper, we use the NLMS algorithm [6] which has much less computational complexity than the recursive least square (RLS) algorithm [10].

Array weight vector obtained after the qth updating is denoted by $\mathbf{W}_q = \left[w_{0,q}\ w_{1,q} \cdots w_{M-1,q}\right]^T$. The initial vector $\mathbf{W}_0$ is set as

$$\mathbf{W}_0 = \left[1 + j0, 0 + j0, \cdots, 0 + j0\right]^T \quad .(11)$$

$\mathbf{W}_q$ is updated using the received pilot as

$$\mathbf{W}'_q = \mathbf{W}_{q-1} + 2\mu e_{q-1}(q \bmod N_c)\frac{\mathbf{R}^*(q \bmod N_c)}{\left\|\mathbf{R}(q \bmod N_c)\right\|^2} \quad ,(12)$$

where μ is the step size parameter and

$$\left\|\mathbf{R}(q \bmod N_c)\right\|^2 = \sum_{m=0}^{M-1}\left|R_m(q \bmod N_c)\right|^2 \quad .(13)$$

We introduce the following constraint to make the array combining output noise power unchanged.

$$\mathbf{W}_q = \frac{\mathbf{W}'_q}{\left\|\mathbf{W}'_q\right\|} \quad .(14)$$

The error signal $e_q(k)$ is the difference between the array output $Y_q(k)$ and the reference $Z_q(k)$. We use the desired signal component after the array combining as the reference signal. $Y_q(k)$ and $Z_q(k)$ are given by

$$\begin{cases} Y_q(k) = \mathbf{R}^T(k)\mathbf{W}_q \\ Z_q(k) = \mathbf{H}_0^T(k)\mathbf{W}_q P_0(k) \end{cases} \quad .(15)$$

e_q(k) is give by

$$e_q(k) = Z_q(k) - Y_q(k) \quad .(16)$$

The average noise power $2\tilde{\sigma}^2$ after the array combining is given by

$$2\tilde{\sigma}^2 = \frac{2N_0N_c}{T_s}\left\|\mathbf{W}_q\right\|^2 \quad .(17)$$

In the proposed scheme, the error signal $e_q(k)$ is the sum of the interference and noise, and the array weight is updated to minimize the squared average of $e_q(k)$. It can be understood from Eq. (17) that the average noise power is constant after the array combining since $\left\|\mathbf{W}_q\right\| = 1$ due to the constraint of Eq. (14). Hence, the proposed scheme works so as to minimize the average interference power.

IV. COMPUTER SIMULATION

The simulation conditions are given in Table 1. We assume QPSK data-modulation, N_c=256 symbols, N_g=32 symbols and U=2. An antenna array with M=2 and an antenna separation of half carrier-wavelength (0.5λ) is considered. As the step size μ gets larger, faster convergence rate is achieved, but the mean square error (MSE) increases. It was found by our preliminary computer simulation that μ=0.06 gives a good tradeoff between the convergence rate and the MSE. Therefore, in the following simulation, μ=0.06 is used. The average received signal-to-interference power ratio (SIR) is set to 0dB (the worst case), and sampling timing and channel estimation are assumed to be ideal.

A. Channel model

In this paper, the distributed scattering model [9] where each propagation path is formed by many local scatterers surrounding a mobile is used. The number of local scatterers is denoted by N_e. It is assumed that L large reflectors are present between the base station and mobile station. A group of N_e element waves created by local scatterers is reflected by the lth (l=0~L−1) large reflector and as a consequence, L propagation paths with different time delays are formed. Fig.3 shows the lth

(l=0~L−1) path propagation model. The angle between the uth user (u=0~U−1) and the antenna axis is denoted by Φ_u and the lth path nominal arrival angle is denoted by $\Phi_{u,l}$. We assume that $\Phi_{u,l}$ (l=0~L−1) is uniformly distributed within a range of ±0.5δϕ from the center of Φ_u. The angle between the nth (n=0~N_e−1) element wave of the lth path and the antenna axis is $\Phi_{u,l}+\alpha_{u,l,n}$. Each element wave suffers different Doppler shift due to the user's movement. When the uth user moves at a speed of v, the nth element wave's Doppler frequency becomes $f_D\cos\beta_{u,l,n}$, where $f_D=v/\lambda$ is the maximal Doppler frequency, λ is the carrier wave length, and $\beta_{u,l,n}$ is the angle between the uth user's moving direction and the nth (n=0~N_e−1) scatterer. The complex path gain $h_{u,m,l}$ of the lth path is expressed as

$$h_{u,m,l}=\frac{1}{\sqrt{N_e}}\sum_{n=0}^{N_e-1}\exp j\left[2\pi\left\{\begin{array}{l}f_D t\cos\beta_{u,l,n}\\+\dfrac{D}{\lambda}\left(\dfrac{M-1}{2}-m\right)\\\cdot\cos(\Phi_{u,l}+\alpha_{u,l,n})\end{array}\right\}+\psi_{u,l,n}\right],\quad(18)$$

where $\psi_{u,l,n}$ represents the random phase of the nth element wave.

In this paper, we assume N_e=32 and that each element wave constituting a path is distributed within a range of ±Δ/2 degrees and the angle spread, Δ=2 degrees, of element waves is assumed. A two-user case (U=2) is considered; the angle between the 0th (1st) user and the antenna axis is assumed to be Φ_0=60° (Φ_1=120°).

TABLE I. SIMULATION CONDITIONS.

Transmitter	Modulation	QPSK
	Pilot	1 block
	Number of FFT points	N_c=256
	GI	N_g=32
Channel	Number of users	U=2
	Number of paths	L=2, 16
	Nominal arrival angles of user 0 and user 1	Φ_0=60°,Φ_1=120°
	Arrival angle spread of L paths	δϕ=2°~30°
	Number of element waves	N_e=32
	Angle spread of element paths	Δ=2°
Receiver	Number of antennas	M=2
	Average SIR	0dB
	Antenna separation	D=0.5λ
	LMS step size parameter	μ=0.06
	Frequency-domain equalization	MMSE
	Channel estimation	Ideal

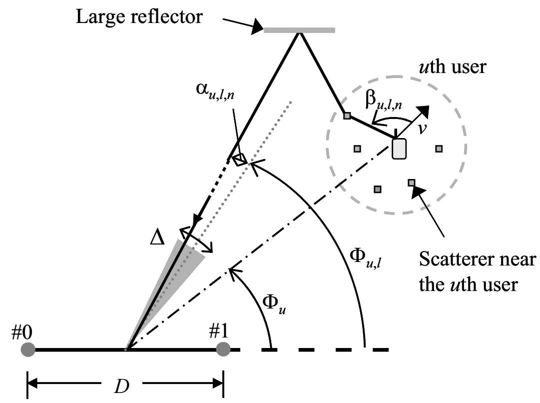


Figure 3. Propagation model.

B. Simulation results

The array weight convergence rate is measured using the average signal-to-interference plus noise power ratio (SINR) after the array combining. The array output $Y_q(k)$ for the qth updated weight $\mathbf{W}_q$ is given, from Eqs (10) and (15), by

$$Y_q(k)=\mathbf{H}_0^T(k)\mathbf{W}_qP_0(k)+\sum_{u=1}^{U-1}\mathbf{I}_u^T(k)\mathbf{W}_q+\mathbf{\Pi}^T(k)\mathbf{W}_q,\quad(19)$$

where $\mathbf{I}_u(k)=\left[I_{u,0}(k)\ I_{u,1}(k)\cdots I_{u,M-1}(k)\right]^T$ is the interference vector. The first term represents the desired signal, the second is the interference and the third is the noise. The instantaneous SINR $\Gamma(k)$ after the array combining is given by

$$\Gamma(k)=\frac{|\mathbf{H}_0^T(k)\mathbf{W}_qP_0(k)|^2}{\sum_{u=1}^{U-1}|\mathbf{I}_u^T(k)\mathbf{W}_q|^2+|\mathbf{\Pi}^T(k)\mathbf{W}_q|^2}.\quad(20)$$

The average SINR after the array combining is plotted in Fig.4 with the angle spread δϕ of L paths as a parameter for L=16 and the average received E_b/N_0=20dB. It can be seen that the array weights converge within 1 block (less than 256 times updating) regardless of δϕ. As δϕ gets larger, the achievable SINR after the weight convergence gets worse; when δϕ=30°, the achievable SINR reduces by about 10dB compared to δϕ=2°. This is because the probability of the arrival angle of the desired user's paths becoming close to that of interfering user increases.

Fig.5 shows the average BER performance with δϕ as a parameter for L=2 and 16. It is seen that the BER performance for L=16 is better than that for L=2, since larger frequency diversity gain can be obtained by MMSE-FDE for L=16 than for L=2. As δϕ gets smaller, the BER performance approaches the single-user's BER performance using adaptive array (for comparison, the BER performance of single-user case without adaptive antenna array is also plotted). When δϕ=2°, the E_b/N_0 degradation from the single user case with adaptive antenna array is as small as 0.8 (2.0) dB at BER=10^{-4} for L=16(2).

V. CONCLUSION

In this paper, frequency-domain adaptive antenna array combined with MMSE-FDE for single-carrier uplink transmission was proposed. The proposed frequency-domain AAA tries to form a beam pattern based on the minimization of the average interference power, therefore the same array weight can be used for all the frequency components. It was shown by computer simulation that although the NLMS algorithm is used for the array weight updating, a very fast weight convergence within one block is achieved. It was also confirmed that the proposed array can obtain the frequency-diversity gain by MMSE-FDE and therefore, it provides a very good BER performance while suppressing the co-channel interference. When the angle spread of paths is $\delta\phi$=2 degrees, the E_b/N_0 degradation from the single user is as small as 0.8 (2.0) dB for L=16 (2).

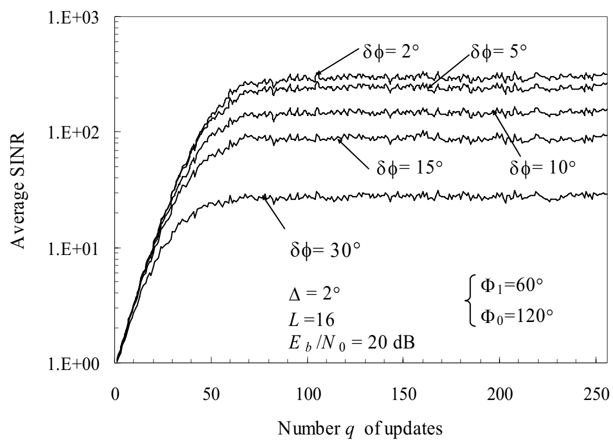
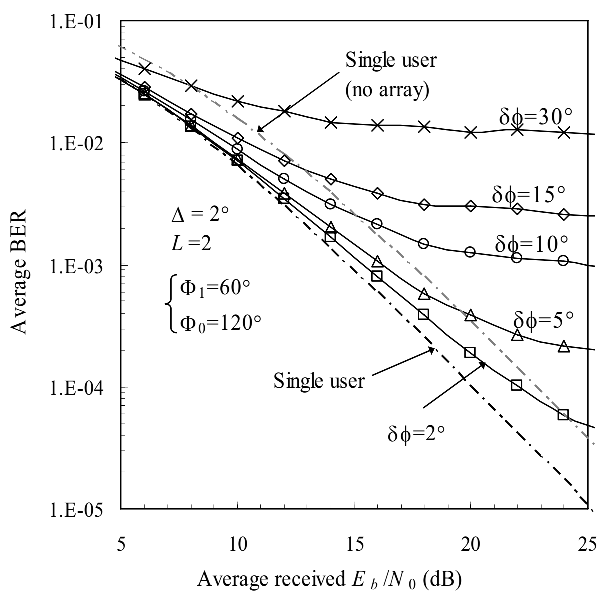
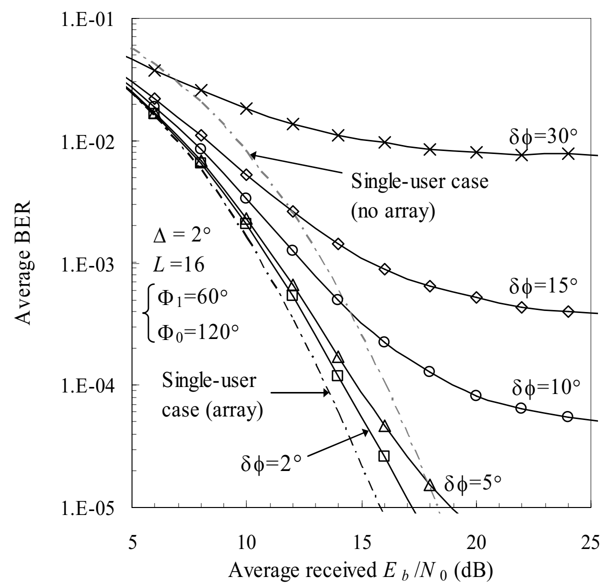


Figure 4. Array output SINR convergence rate.



(a) L=2



(b) L=16

Figure 5. Average BER performance.

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