A New Spatio-Temporal Equalization Method Using Estimated Channel Impulse Response

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Abstract - This paper proposes a new spatio-temporal equalization method, which utilizes an adaptive antenna array and a decision feedback equalizer (DFE) simultaneously. In the design of a spatio-temporal equalizer, how to split equalization functionality into spatial equalizer and temporal equalizer plays an important role to achieve effective equalization. Since the performance of spatial equalizer and temporal equalizer depends on channel conditions, such as direction of arrival (DoA) patterns and maximum time delay of incoming signals, it is essential to adaptively split the functionality into the spatial and temporal equalizer, according to the channel condition. The proposed equalizer adaptively changes its beamforming method depending on estimated channel impulse response, therefore, it can achieve high performance in various channels. In this paper, the performance of the proposed system in Rayleigh fading channels, Rician fading channels and static multipath channels, in terms of bit error rate (BER), is evaluated in comparison with the performance of spatial equalizer and temporal equalizer.

I. Introduction

Spatio-temporal equalization is a technique which utilizes both spatial and temporal information of received signal to compensate intersymbol interference due to multipath fading. In the design of the spatio-temporal equalizer, how to split equalization functionality into spatial equalizer and temporal equalizer plays an important role to achieve effective equalization. Taking into account the computational complexity, the spatial equalizer is superior to the temporal equalizer, because the complexity of the spatial equalizer does not depend on maximum time delay of incoming signals, whereas the temporal equalizer does. In this sense, the spatial equalizer is more suitable for high speed wireless communications systems than the temporal equalizer. On the other hand, taking into consideration the sensitivity to direction of arrival (DoA) patterns, the temporal equalizer is more preferable than the spatial equalizer because the temporal equalizer does not care the difference of DoA patterns, whereas the performance of the spatial equalizer mainly depends on DoA patterns. It follows from what has been mentioned that it is essential to adaptively split equalization functionality into the spatial and temporal equalizer, according to channel conditions.

We have proposed a spatial equalization (beamforming) method for indoor high speed wireless multimedia communications systems[1],[2], where the beamformer calculates and adjusts the weights of adaptive array sensors using estimated channel impulse response. Furthermore, we have extended the beamforming method to a spatio-temporal equalization method[3], where the fundamental criterion is 'delayed incoming signals with large time delays should be canceled by the spatial equalizer'. The spatio-temporal equalizer can achieve fairly good performance, however, there may be room for improvement, because the equalizer employs fixed equalization functionality assignment.

In this paper, we propose a new spatio-temporal equalization method, which utilizes an adaptive antenna array and a decision feedback equalizer (DFE) simultaneously. What is worthy of special mention about the equalizer is that it selects a suitable beamforming method for DoA pattern using estimated channel impulse response. This means that the equalizer can adaptively change the way of assignment of equalization functionality. Moreover, the selection of beamforming method is based only on the change of reference signal for the beam-weights calculation, therefore, the proposed equalizer requires no changes in the hardware even when some modifications in functionality assignment algorithm are needed.

We also show the bit error rate (BER) performance of the proposed spatio-temporal equalizer in various channels, such as Rayleigh fading channels, Rician fading channels and static multipath channels and compare the performance among a temporal equalizer (DFE), a spatial equalizer (our already-proposed beamformer), and the proposed spatio-temporal equalizer.

II. System Configuration

Assume that the proposed spatio-temporal equalizer is applied to the down link of a wireless local area network (LAN) system, where a base station with an omnidirectional antenna communicates with n half-fixed terminals each having an adaptive antenna array and DFE. Fig. 1 shows the transmitter/receiver structure. In our method, we prepare two types of channels; the traffic channel and the pilot channel. The traffic channel is used to convey information signals, and the pilot channel is for the pilot signal which

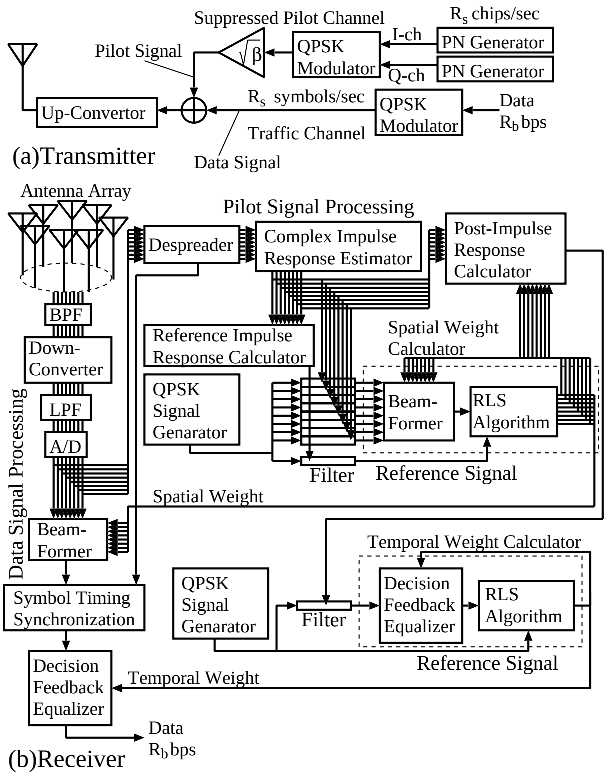

Fig. 1. Transmitter/Receiver Structure

is used to estimate the channel impulse response.

In the transmitter, data sequence (R_b [bits/sec]) is first converted into QPSK waveform ($R_s(=R_b/2)$ [symbols/sec]) for the traffic channel, on the other hand, the pilot signal is generated by the pseudo noise (PN) sequence generator for the pilot channel, which has the same frequency band width as the traffic channel. The power of the pilot signal is suppressed enough before the addition to the data signal. If we use a S_{pn}-stage maximum length shift resister (M) sequence as a PN sequence, the period of the PN sequence becomes $P\ (= 2^{S_{pn}} - 1)$ symbols long. In addition, different sequences are used for the inphase and quadrature channels (I and Q-ch).

Let $s_B(t)$, $s_{Tr}(t)$, and $s_{Pi}(t)$ denote the baseband transmitted signal, the data signal, and the pilot signal, respectively. They can be written as

$$s_B(t) = s_{Tr}(t) + \sqrt{\beta} \cdot s_{Pi}(t), \quad (1)$$

$$s_{Tr}(t) = \sum_{k=-\infty}^{\infty} d_k \delta(t - kT_s), \quad (2)$$

$$s_{Pi}(t) = \sum_{l=-\infty}^{\infty} \sum_{k=0}^{M-1} m_k \delta(t - (lM + k)T_s), \quad (3)$$

$$d_k = d_{I_k} + j d_{Q_k}, \quad (4)$$

$$m_k = m_{I_k} + j m_{Q_k}, \quad (5)$$

where d_{I_k} and d_{Q_k} denote the kth symbol of the inphase and quadrature components in the data signal, respectively. T_s, β, $\delta(y)$, and M denote the time duration of one symbol, the power suppression ratio of the pilot channel, the Dirac's delta function, and the period of M sequence, respectively. Moreover, m_{I_k} and m_{Q_k} express the inphase and quadrature components of the pilot signal at the kth symbol, respectively.

The baseband signal passes through the low pass filter (LPF), i.e., emission filter, and then is transmitted from the omnidirectional antenna after up-conversion. The transmitted signal $s_T(t)$ will be

$$s_T(t) = Re[z_0(t) \exp(j2\pi f_c t)], \quad (6)$$

$$z_0(t) = s_B(t) * h_B(t), \quad (7)$$

where $h_B(t)$, f_c, and $z_0(t)$ denote the impulse response of the LPF, the carrier frequency, and the band-limited baseband signal, respectively.

In the receiver, the incoming signal is received by the antenna array which consists of N_{ary} sensors, where the sensor spacing is the half of carrier wavelength. After then, the received signal undergoes the band pass filter (BPF), down-converter, LPF (matched filter), and A/D converter. The BPF is used for the suppression of the adjacent channel interference and noise as well as for the extraction of the spectrum around the desired signal. Moreover, the A/D converter operates with the sampling rate of N_{smp} times the symbol rate. After the A/D conversion, the received signal is processed in the data signal processing section and in the pilot signal processing section independently.

In the data signal processing section, the outputs from the matched filters are multiplied by the weights of the beamformer which are calculated in the pilot signal processing section. After symbol timing synchronization and equalization with DFE which has N_{ftap} taps in the feedforward filter and N_{btap} taps in the feedback filter, the data are recovered.

In the pilot signal processing section, the complex instantaneous channel impulse response at each sensor is first estimated by despreading the received pilot signal. Moreover, N_{pnseq} responses so obtained are coherently added to suppress the traffic channel sufficiently. Since the power of the pilot signal is suppressed by the factor of β before the transmission, the effective processing gain of the pilot channel G_{total} will be

$$G_{total} = P \times N_{pnseq} \beta. \quad (8)$$

Then, using the estimated channel impulse response, the receiver selects a suitable beamforming method for the channel condition. Details of the selection of beamforming method, in other words reference signal generation, are discussed in section III. Next, a QPSK signal is generated in the receiver and is fed into the filter whose tap coefficients are the same as the estimated channel impulse response, and as a result, the output of the filter becomes a pseudo-received signal. On the other hand, the QPSK signal is also fed into the filter whose tap weights are the same as the impulse response for the reference signal generation. Using the output of this filter as a

reference signal and the pseudo-received signal, the weights of sensors are calculated by using recursive least square (RLS) algorithm[4].
Now that the estimated channel impulse response and the weights of the adaptive antenna array sensors are available, we can calculate the tap weights of the DFE by using RLS algorithm in the same manner as the beam-weights calculation.

III. Beamforming Method

A. Channel Response Estimation

In the pilot signal processing section, we first estimate the channel impulse response. Let $g_l(t)$ denote the output of the matched filter at the lth antenna element. In the proposed system, since the observation window width of the channel impulse response is equal to P symbols, the estimated channel impulse response at the kth estimation window can be written as

$$\hat{h}_l^k(\tau) = \sum_{i=0}^{N_{smp}\times P-1} g_l(t_k + i\frac{T_s}{N_{smp}})\delta(\tau - i\frac{T_s}{N_{smp}}), \quad (0 \le \tau \le P\times T_s) \qquad (9)$$

where $\hat{(\cdot)}$ denotes the estimation of $(\cdot)$.
As we mentioned in section II, the channel impulse response is estimated using $P \times N_{pnseq}$ pilot signals, therefore, the estimated channel impulse response at the lth antenna element $\hat{h}_l$ will be obtained by averaging out $\hat{h}_l^k(\tau)$, namely,

$$\hat{h}_l(\tau) = \frac{1}{N_{pnseq}} \sum_{k=1}^{N_{pnseq}} \hat{h}_l^k(\tau). \qquad (10)$$

Next, we search for a path with the maximum power. In other words, defining

$$\sigma(\tau) = \sum_{l=1}^{N_{ary}} |\hat{h}_l(\tau)|^2, \qquad (11)$$

as the total power at τ, we search for $\tau = \tau_{max}$ such that $\sigma(\tau)$ is maximal. Using τ_{max}, we can estimate the channel impulse response at the lth antenna element including the path with the maximum power as

$$f_l(k) = \hat{h}_l(\tau_{max} + kT), \quad (1 \le l \le N_{ary}), \qquad (12)$$

which has a non-zero value for k satisfying $0 \le \tau_{max} + kT_s \le P \times T_s$ and zeros otherwise.
The impulse response $f_l(k)$ is used to generate the pseudo-received signal at the lth antenna element $x'_l(k)$:

$$x'_l(k) = d'(k) * f_l(k) + n'_l(k), \qquad (13)$$

where $*$ denotes the convolution, $d'(k)$ is the QPSK signal generated in the receiver, and $n'_l(k)$ is also the generated noise, whose power is equal to that of the noise in the channel.

B. Reference Signal Calculation

In the proposed system, two types of reference signals are prepared: reference signal to capture only the path with the maximum power(*ref.1*) and reference signal to capture not only the path with the maximum power but also delayed signals(*ref.2*). Both reference signals are generated by the impulse response $f_{ref}(k)$:

$$f_{ref}(k) = \frac{1}{N_{ary}}(\sum_{l=1}^{N_{ary}} |f_l(k)|)\frac{f_1(k)}{|f_1(k)|}, \qquad (14)$$

though $f_{ref}(k)$ is equal to zero except for $k = 0$ for *ref.1* generation, whereas $f_{ref}(k)$ is zero for $k < 0$ or $k > N_{feedback}$ for *ref.2* generation. Note that, for *ref.2*, incoming signals whose delay time exceed the length of feedback filter in the DFE are canceled at the spatial equalizer.
Using $f_{ref}(k)$, the reference signal $x'_{ref}(k)$ can be written as

$$x'_{ref}(k) = d'(k) * f_{ref}(k). \qquad (15)$$

C. Reference Signal Selection

Here, we mention how to select the reference signal. Now we consider two types of radio wave propagation model:

• *Case 1: no simultaneous propagation path*

In this propagation model, only one signal arrives at the antenna array in a moment. Therefore, we can estimate DoA of incoming signals from the phase difference of the estimated channel impulse response at each sensor. Accordingly, the receiver can tell whether the spatial equalizer can cancel the delayed signal or not, namely, the receiver select *ref.2* when the primary signal is in the same direction (within θ_{near}) as the delayed signal and select *ref.1* otherwise.

• *Case 2: Some simultaneous propagation paths*

In this case, the received signal no longer has DoA information, therefore, we can not use the same approach as in *Case 1*. However, we have shown in [1] that our already proposed beamformer works well even when the primary signal and the delayed signal are in the same direction, if there are some incoming signals at the same timing as the primary signal or the delayed signal. This means that we can employ *ref.1* in *Case 2* regardless of DoA patterns.
Then, what becomes a question is how to tell *Case 1* from *Case 2*. In *Case 2*, we can reasonably expect that the estimated channel response at each sensor will vary in the power because of a standing wave and such a phenomenon does not occur in *Case 1*. Therefore, defining the normalized variation of the estimated channel response among sensors $P_{vari}(k)$ as

$$P_{vari}(k) = \frac{1}{N_{ary}} \sum_{l=1}^{N_{ary}} \frac{||f_l(k)| - |f_1(k)||^2}{|f_1(k)|^2}, \qquad (16)$$

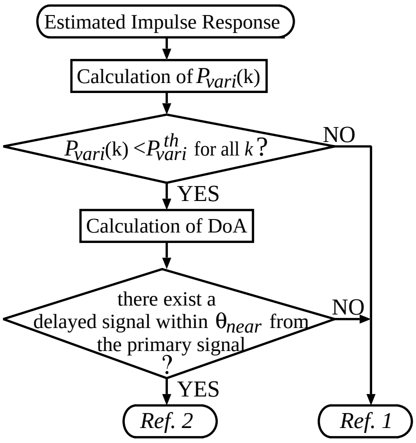

Fig. 2. Reference Signal Selection Algorithm

TABLE I
TAPPED-DELAY-LINE PARAMETERS

tap	Channel Model A		Channel Model B	
	rel. delay [symbol]	avg. power [dB]	rel. delay [symbol]	avg. power [dB]
1	0	0	0	0
2	0.5	-3.0	1.0	-3.6
3	1.0	-10.0	2.0	-7.2
4	1.75	-18.0	3.0	-10.8
5	3.0	-26.0	5.0	-18.0
6	3.0	-32.0	7.0	-25.2

TABLE II
PARAMETERS

	proposed	spatial	temporal
Num. of sensors N_{ary}	8	8	1
Num. of feedfoward taps	5	-	10
Num. of feedback taps	4	-	9
Symbol rate R_s	100[Msymbols/sec]		
Carrier frequency f_c	60[GHz]		
Roll-off factor α	0.5		
Pilot suppression ratio β	0.04		
M-sequence Period P	255 [symbols]		
Num. of additions N_{pnsec}	16		
Oversampling factor N_{smp}	4		
RLS Repetitions	50		

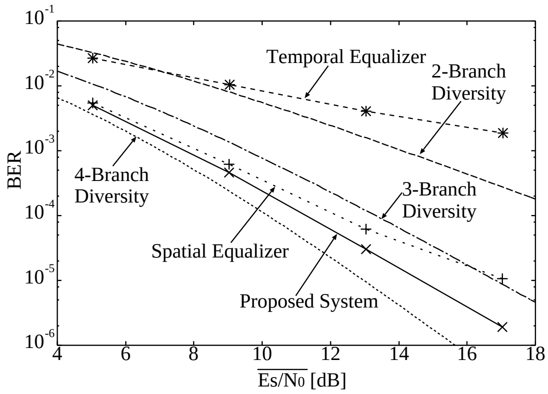

Fig. 3. BER Performance in Channel Model A (Rayleigh)

we judge the channel has no simultaneous path (*Case 1*), if $P_{vari}(k)$ does not exceed a threshold P^{th}_{vari} for all k.
At the end of this section, we summarize the reference signal selection algorithm in Fig.2.

IV. Computer Simulation

Computer simulations are conducted to evaluate the performance of the proposed spatio-temporal equalizer in comparison with a temporal equalizer (DFE[5]), and a spatial equalizer (our already-proposed beamformer[1]).

A. Channel Model

In these simulations, as an example of indoor channel environment, we use "Indoor Office Test Environment Tapped-Delay-Line Parameters" defined in [6]. The parameters are given by Table I. For each channel model, we further assume three types of channels: a Rayleigh fading channel, a Rician fading channel and a static multipath channel. In all the channel models, DoA of each incoming signal is randomly determined and it changes every frame timing. Taking account of an indoor environment, we have chosen the Doppler spectrum of flat and the maximum Doppler shift of 150Hz.

B. Parameter

Parameters used in all the computer simulations are summarized in Table II. The proposed system and the spatial equalizer employ an circular antenna array with eight sensors. On the other hand, the temporal equalizer employs an omnidirectional antenna, therefore, in the following figures, the curves of the BER performance of the temporal equalizer will be shifted by 9[dB]. In the proposed system, thresholds of the DoA θ_{near} and the normalized variation P^{th}_{vari} were chosen to be 10.0[deg] and 0.029, respectively.

C. Bit Error Rate Performance

Fig.3-5 show the BER performance versus the ratio of the average energy per symbol to the noise power density($\overline{E_s/N_0}$) for the Rayleigh fading channel, Rician fading channel and static multipath channel, respectively, for the Channel Model A. Fig.6-8 also show the BER performance versus the $\overline{E_s/N_0}$ for the Rayleigh fading channel, Rician fading channel and static multipath channel, respectively, for the Channel Model B. Though the proposed system is not an antenna diversity system because of the small sensor spacing, it considered to be meaningful to compare the results with the theoretical bit error rate of diversity reception. Therefore, the theoretical BERs are also plotted in Fig.3 and Fig.6.
In all the figures, the proposed system can achieve the best performance among the three equalization methods. Moreover, the proposed system can achieve the gain of 3-branch maximal-ratio combining diversity in Fig.3 and the gain of 2-branch in Fig.6. From the results, it can be concluded that the proposed system can attain high performance in various channels and

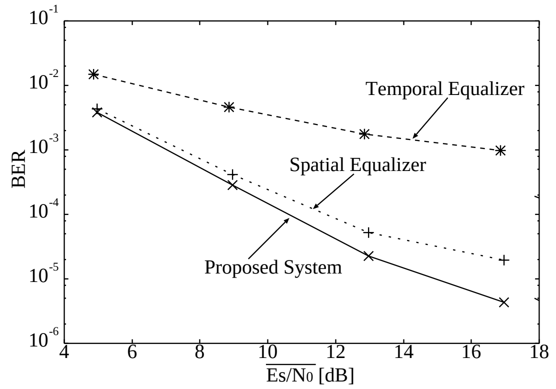

Fig. 4. BER Performance in Channel Model A (Rician)

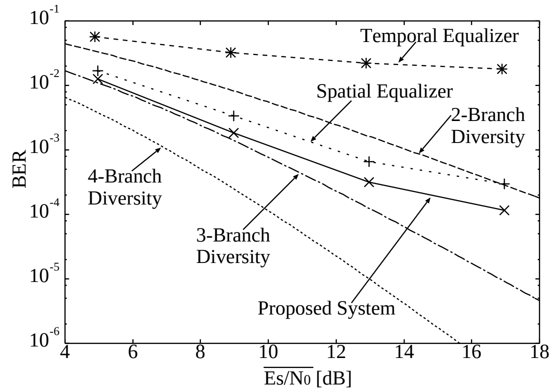

Fig. 6. BER Performance in Channel Model B (Rayleigh)

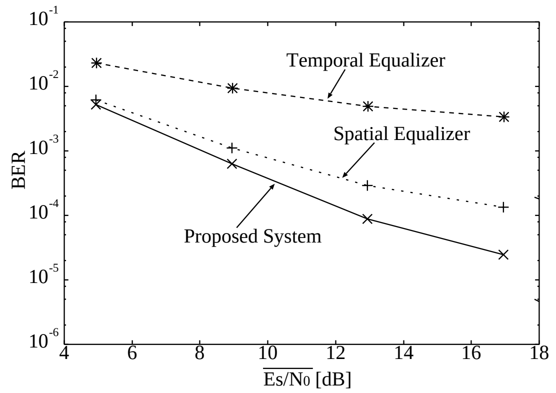

Fig. 5. BER Performance in Channel Model A (static)

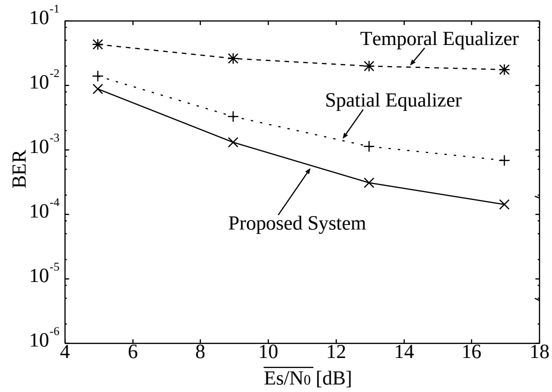

Fig. 7. BER Performance in Channel Model B (Rician)

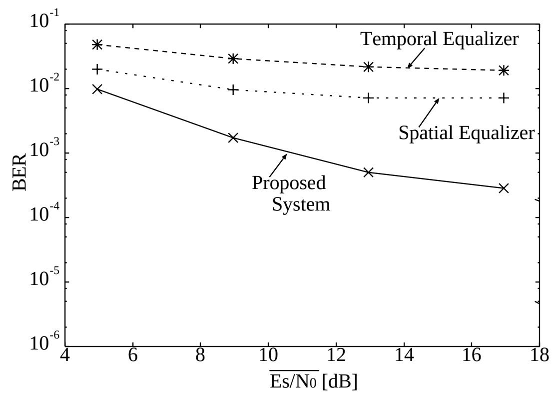

Fig. 8. BER Performance in Channel Model B (static)

robust to change of propagation condition.

V. Conclusion

In this paper, we have proposed a new spatio-temporal equalization method using estimated channel response and evaluated its performance comparing with the spatial equalizer and the temporal equalizer. What is especially important about the proposed equalizer is that it can select a suitable beamforming method depending on the channel condition. In addition, the selection is based only on the change of reference signal for beam-weights calculation, therefore, it requires no changes in hardware even when some modifications in beamforming method selection algorithm are needed.

We have shown the BER performance in Rayleigh fading channels, Rician fading channels, and static multipath channels. From all the results, it can be concluded that the proposed system can achieve the best and stable performance among the three equalization methods in all the channel models.

With our approach, we could easily develop the proposed system to work in more intelligent manner. The investigation of the equalization functionality assignment algorithm is in progress.

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