# A NEW SPATIO-TEMPORAL EQUALIZATION METHOD WITHOUT DOA INFORMATION

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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. For effective spatio-temporal equalization, how to assign the role to spatial equalization part and temporal equalization part is really important. One of the answers which we have given is 'incoming signals with larger time delays should be cancelled at the spatial equalization part.' The weights of both adaptive antenna array elements and taps of DFE are calculated only using the estimated channel impulse response, therefore, it requires no information on direction of arrival (DoA). We show the performance of the proposed system in multipath fading channels often encountered in indoor wireless environments, and discuss the attainable bit error rate (BER) and the computational complexity in comparison with other equalization methods, such as spatial equalization and temporal equalization.

## I. INTRODUCTION

Spatio-temporal equalization is a technique which utilizes both spatial and temporal information of the received signal to compensate intersymbol interference due to multipath fading. What is important to realize effective spatio-temporal equalization is how to assign the role to spatial equalization part and temporal equalization part. In this paper, the fundamental strategy is based on 'incoming signals with larger time delays should be cancelled at the spatial equalization part.' This is because the performance of the spatial equalizer does not depend on the delay time, on the other hand, the computational complexity of temporal processing largely depends on the maximum length of the time delay.

We have been proposing a spatial equalization (beamforming) method for indoor high speed wireless multimedia communications systems[1], where the beamformer calculates and adjusts the weights of adaptive array elements using suppressed spread spectrum (SS) pilot signals. The SS pilot signals are parallelly transmitted with data signal with the power suppressed enough, therefore, the receiver can handle only the pilot channel, independent of the data channel. After estimating the instantaneous impulse response at each antenna element with the suppressed SS signals, the weights are adjusted so as to maximize the signal to noise plus interference ratio (SNIR) before signal demodulation process. We have clearly showed that the weights of adaptive antenna array elements can be controlled by the estimated channel impulse response and that the proposed beamformer can achieve a better bit error rate (BER) performance than a constant modulus algorithm (CMA) based beamformer. In addition, we have discussed a blind approach requiring no pilot signal[2].

On the other hand, temporal equalization methods such as decision feedback equalizer (DFE) and maximum-likelihood sequence estimation (MLSE) generally equalize distorted channel, employing the estimated channel impulse response. In this sense, it could be easier to combine our proposed beamformer with a temporal equalization, because the estimated channel impulse response is employed in both spatial and temporal equalization part.

In this paper, we discuss a new spatio-temporal equalization method employing the channel impulse response estimation. We show the bit error rate performance for the proposed spatio-temporal equalizer in multipath fading channels often encountered in indoor wireless environments. We also compare the performance among a temporal equalizer (DFE), a spatial equalizer (our already-proposed beamformer), and the spatio-temporal equalizer. In addition, we evaluate the computational complexity required in the spatio-temporal processing.

## II. SYSTEM CONFIGURATION

It is assumed that the proposed spatio-temporal equalizer is applied to down link of wireless LAN (Local Area Network) 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 pilot channel is used to estimate the channel impulse response for the weights (of both spatial and temporal equalizers) calculation.

In the transmitter, data sequence (200Mbit/sec) is first converted into QPSK waveform (100Msymbol/sec) as the traffic channel. The pilot channel is generated by the pseudo noise (PN) sequence gener-

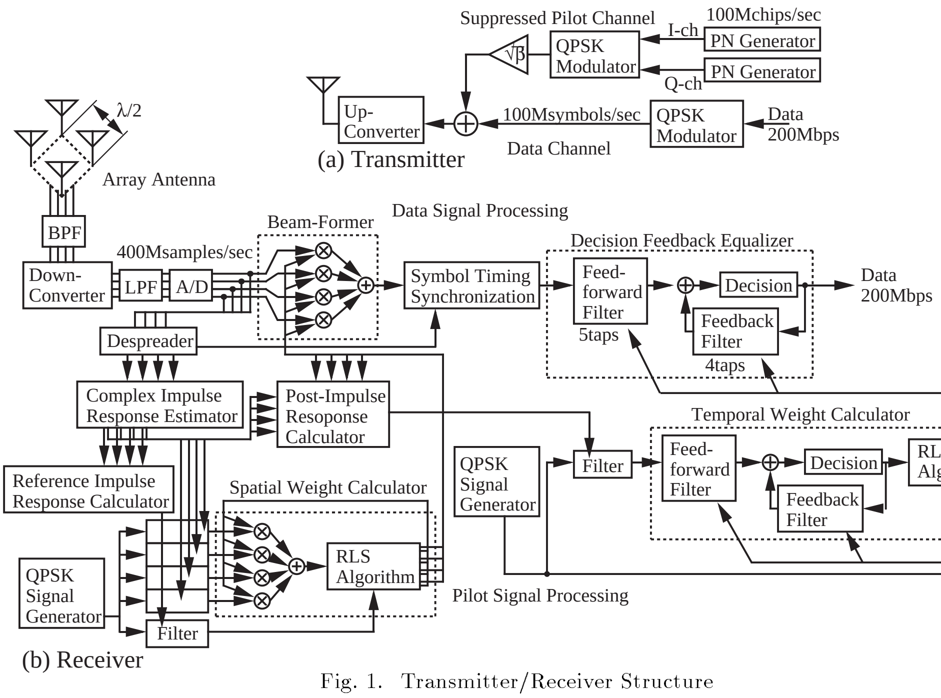


Fig. 1. Transmitter/Receiver Structure

ator and has the same frequency band as the traffic channel. The power of the pilot channel is suppressed enough before the addition to the traffic channel. In this paper, since we use the 8-stage maximum length shift resister (M) sequences as PN sequences, and the two different PN sequences with period of 255 are used in the inphase and quadrature channels, respectively.

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

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

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

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

$$d_k = d_{I_k} + j d_{Q_k} \tag{4}$$

$$m_k = m_{I_k} + j m_{Q_k}, \tag{5}$$

where $d_{I_k}$ and $d_{Q_k}$ denote the $k$th symbol of the inphase and the quadrature components in the traffic channel, respectively. $T_s$, $\beta$, $\delta(y)$, and $M$ denote the time duration of one symbol (=10ns), the power suppression ratio of the pilot channel (= 0.04), Dirac's delta function, and the period of M sequence, respectively. $m_{I_k}$ and $m_{Q_k}$ express the inphase and quadrature components of the pilot channel at the $k$th symbol, respectively.

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

$$s_T(t) = Re[z_0(t) \exp(j 2\pi f_c t)] \tag{6}$$

$$z_0(t) = s_B(t) * h_B(t), \tag{7}$$

where $h_B(t)$, $f_c$, and $z_0(t)$ denote the impulse response of the LPF, the carrier frequency (=60 GHz), and the band-limited baseband signal, respectively.

In the receiver, the incoming signal is received by the antenna array which consists of $N_{ary}(=4)$ sensors arranged in the position of the vertex of a square, where the sensor spacing is the half of carrier wavelength. The received signal undergoes the 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. Four times over sampling is employed in the A/D converter. Next, the received signal is processed in the traffic channel processing part and in the weight calculation part independently.

In the traffic channel processing part, the outputs from the matched filters are multiplied by the weights of the beamformer which are calculated in the weight calculation part. After symbol timing synchronization and equalization with DFE which has $N_{tap}$(=9) taps (5taps in the feedforward filter and 4taps in the feedback filter), the data are recovered.

In the weight calculation part, the complex instantaneous channel impulse response at each antenna element is first estimated by despreading the pilot channel. Since $N_{pnseq}$(=16) PN sequences are coherently added, the processing gain in the pilot chan-

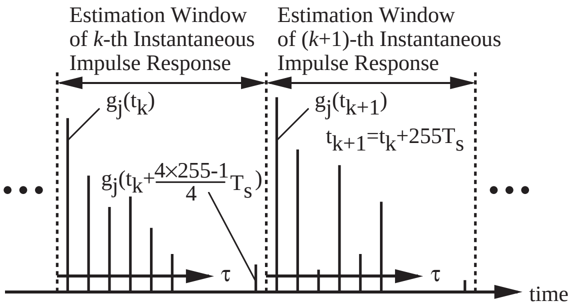


Fig. 2. Estimated Instantaneous Impulse Response

nel becomes $255 \times N_{pnseq}$. The estimated channel impulse response is used to calculate the impulse response for the reference signal generation. Details of the 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. As a result, the output of the filter becomes a pseudo-received signal. 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 the reference signal, the weights of antenna elements are calculated by the RLS algorithm with the pseudo-received signal.

Now that the estimated channel impulse response and the weights of the adaptive antenna array elements are available, we can calculate the tap weights of the DFE by the RLS algorithm in the same manner as the beam-weights calculation.

## III. BEAM-WEIGHT CALCULATION

In the weight calculation part, we first calculate the channel impulse response. Let $g_j(t)$ denote the output of the matched filter at the $j$th antenna element. In our proposed system, since the time window width of the channel impulse response observation is equal to 255 symbols, the estimated channel impulse response at the $k$th estimation window can be written as

$$\hat{h}_j^k(\tau) = \sum_{i=0}^{4\times 255-1} g_j(t_k + i\frac{T_s}{4})\delta(\tau - i\frac{T_s}{4}), \quad (0 \le \tau \le 255T_s) \qquad (8)$$

where $\hat{(\cdot)}$ denotes the estimation of $(\cdot)$. Fig.2 shows an example of the estimated impulse response. As we mentioned in section II, the channel impulse response is estimated using $255 \times N_{pnseq}$ pilot signals, therefore, the estimated channel impulse response at the $j$th antenna element $\hat{h}_j$ will be obtained by averaging out $\hat{h}_j^k(\tau)$, namely,

$$\hat{h}_j(\tau) = \frac{1}{N_{pnseq}} \sum_{k=1}^{N_{pnseq}} \hat{h}_j^k(\tau). \qquad (9)$$

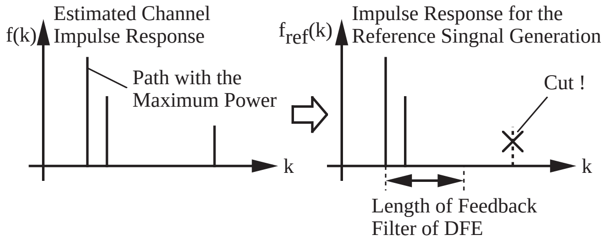


Fig. 3. Reference Impulse Response

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

$$\sigma(\tau) = \sum_{j=1}^{N_{ary}} |\hat{h}_j(\tau)|^2 \qquad (10)$$

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

$$f_j(k) = \hat{h}_j(\tau_{max} + kT), \quad (1 \le j \le 4), \qquad (11)$$

which has a non-zero value for $k$ satisfying $0 \le \tau_{max} + kT_s \le 255T_s$ and zeros otherwise. The impulse response $f_j(k)$ is used to generate the pseudo-received signal at the $j$th antenna element $x'_j(k)$:

$$x'_j(k) = d'(k) * f_j(k) + n'_j(k), \qquad (12)$$

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

The impulse response for the reference signal generation $f_{ref}(k)$ can be wirtten as

$$f_{ref}(k) = \frac{1}{N_{ary}}(\sum_{j=1}^{N_{ary}} |f_j(k)|)\frac{f_1(k)}{|f_1(k)|}, \quad (0 \le k \le N_{feedback}) \qquad (13)$$

where $N_{feedback}$ denotes the length of the feedback filter in the DFE, and $f_{ref}(k)$ is equal to zero for $k > N_{feedback}$. In the proposed system, the incoming signal whose delay time exceeds the length of feedback filter in the DFE is canceled at the spatial equalization part. Fig.3 shows how to derive $f_{ref}(k)$ from $f_j(k)$.

The reference signal $x'_{ref}(k)$ can be written as

$$x'_{ref}(k) = d'(k) * f_{ref}(k). \qquad (14)$$

With this reference signal $x'_{ref}(k)$ and the pseudo-received signal $x'_j(k)$, the weights of beamformer are calculated by the RLS algorithm.

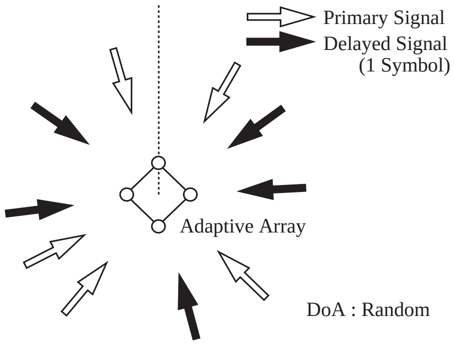


Fig. 4. Channel Model A (frequency selective fading channel)

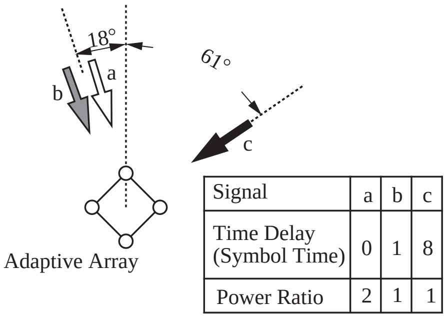


Fig. 5. Channel Model B (static 3-ray multipath channel)

# IV. COMPUTER SIMULATION

## A. Channel Model

Fig.4 shows the frequency selective fading channel model discussed, where there are 5 preceding waves and 5 one-symbol delayed waves. The power ratio of these 10 incoming signals are the same. 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 shift of 150Hz.

Fig.5 shows the static 3-ray multipath channel model. There is a preceding wave in the same direction as a two-symbol delayed wave, and there is one more incoming wave which has a large (eight-symbol) delay time. we discuss the attained performance based on the channel model B among a temporal equalizer (DFE), the spatial equalizer (our already-proposed beamformer), and the proposed spatio-temporal equalizer.

## B. Parameter

In the proposed system, we adopt a root Nyquist filter with roll off factor of 0.5 as the LPF in the transmitter and receiver. The number of repetitions of the RLS algorithm in the both spatial equalization part ($N_{cal\_S}$) and temporal equalization part ($N_{cal\_T}$) are chosen to be 50.

On the other hand, the temporal equalizer employs an omnidirectional antenna and a DFE which has 19taps (10taps in the feedforward filter and 9 in the

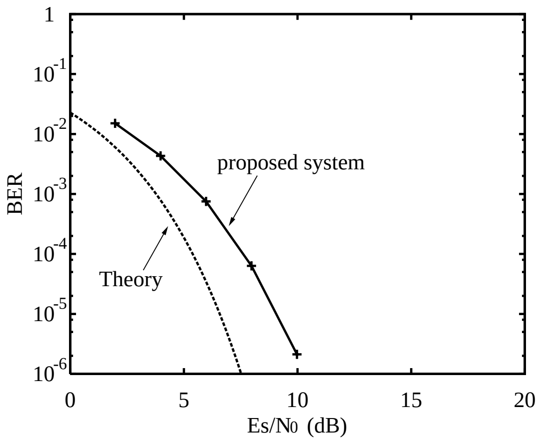


Fig. 6. Bit Error Rate Performance in AWGN Channel

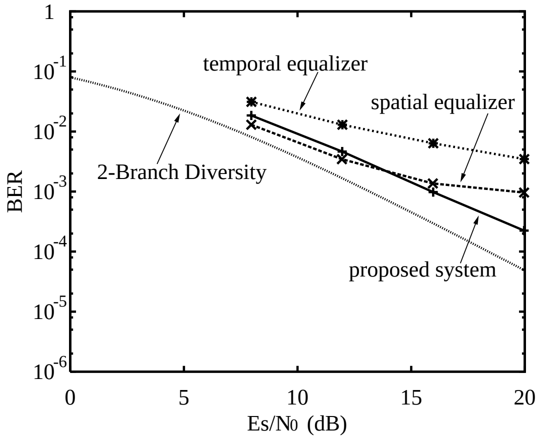


Fig. 7. Bit Error Rate Performance in Channel Model A

feedback filter). The same root Nyquist filters are employed in the temporal equalizer and the spatial equalizer.

## C. Bit Error Rate Performance

Fig.6 shows the bit error rate (BER) performance versus the $E_s/N_0$ in an additive white Gaussian noise (AWGN) channel. We can recognize about 2dB degradation from the theoretical line. This may be because of the insertion of the pilot channel, and the error in the channel impulse response estimation or in the spatial and temporal weights calculation.

Fig.7 shows the BER performance in the channel model A. The BER performance of the spatial equalizer (our already-proposed beamformer) and the temporal equalizer is also plotted in the same figure. Since the temporal equalizer employs an omnidirectional antenna, we have shifted the $E_s/N_0$ of the temporal equalizer by 6dB. The BER of the temporal equalizer improves gradually as $E_s/N_0$ increases, but the performance is the worst among the three equalizers. The spatial equalizer can achieve the best performance for $E_s/N_0 < 14$dB, however, we can recognize the BER floor for $E_s/N_0 > 15$dB. This is be-

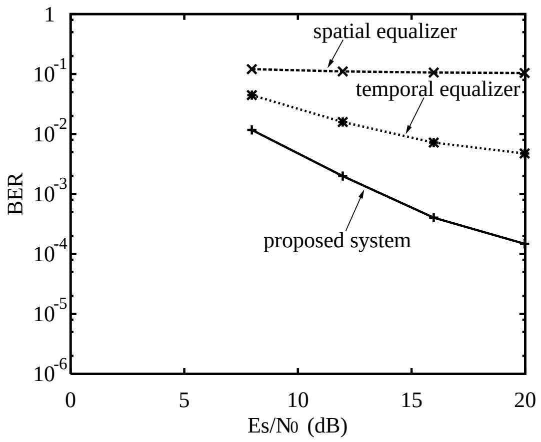

Fig. 8. Bit Error Rate Performance in Channel Model B

cause the burst errors occur in certain DoA patterns. The proposed system can achieve the gain of 2-branch maximal-ratio combining diversity. This means that the equalization of the proposed system does not care DoA patterns.

Fig.8 shows the BER performance in the channel model B. As expected in the BER performance in the channel model A, the equalization of the spatial equalizer breaks down in this channel model. This is because our already-proposed beamformer forms its directional beam as capturing the path with the maximum power, therefore, it try to catch the signal a which is collapsed by the signal b. The performance of the temporal equalizer also improves gradually as $E_s/N_0$ increases in this channel model. However, the proposed system can achieve much better performance. This means that the spatial equalization part in the proposed system works well, that is, it successfully captures the path with the maximum power (signal a) including the delayed signal with no large delay time (signal b). All the temporal equalization part in the proposed system have to do is to cancel the signal b.

### *D. Computational Cost*

We show the number of multiplications in the weight calculation algorithm as a computational cost. In the case of the proposed system, the number of multiplication in the weight calculation algorithm is

$$\begin{aligned} &N_{cal\_S} \times (4N_{ary}^3 + 12N_{ary}^2 + 22N_{ary} + 1) \\ &\quad + N_{cal\_T} \times (4N_{tap}^3 + 12N_{tap}^2 + 22N_{tap} + 1) \\ &\qquad = 231'200 \ (times). \end{aligned}$$

In the case of the spatial equalizer,

$$\begin{aligned} &N_{cal\_S} \times (4N_{ary}^3 + 12N_{ary}^2 + 22N_{ary} + 1) \\ &\qquad = 26'850 \ (times). \end{aligned}$$

And in the case of the temporal equalizer (DFE),

$$\begin{aligned} &N_{cal\_T} \times (4N_{tap}^3 + 12N_{tap}^2 + 22N_{tap} + 1) \\ &\qquad = 1'609'350 \ (times). \end{aligned}$$

The spatial equalizer requires the least computational complexity, but it suffers from break down of the equalization in certain DoA patterns. The computational complexity of the temporal equalizer depends on the maximum time delay of the incoming signal which we must equalize. Therefore, the temporal equalizer may be not suited for high speed communications. The proposed spatio-temporal equalizer not only can achieve an excellent BER performance but also requires a relatively low computational cost.

## V. CONCLUSION

In this paper, we have proposed a new spatio-temporal equalization method without DoA information. The proposed equalizer only uses estimated channel impulse response to adjust the weights of adaptive antenna array elements and the tap weights of the DFE. Here, our fundamental criterion is 'delayed incoming waves with large time delays should be canceled by the spatial equalization part'.

We have shown the attainable bit error rate performance in a AWGN channel, a frequency selective fading channel, and a static 3-ray multipath channel. We have also shown the computational complexity in terms of the number of multiplications in the total weight calculation. The proposed system can achieve the best and stable performance with a relatively low computational complexity.

We have clearly shown that our approach of controlling the adaptive antenna array by the estimated channel impulse response is really suited for spatio-temporal equalization. Many blind channel identification algorithms are now available[3], [4], [5], therefore, our proposed spatio-temporal equalization method may be able to be developed to work in a blind manner by employing such algorithms.

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