Optimum Spatial-Temporal Equalization for Diversity Receiving Systems with Co-channel Interference

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Abstract— **In this paper, we investigate optimum spatial-temporal equalization for diversity receiving systems with co-channel interferences. We first present the structure of the optimum spatial-temporal decision-feedback equalizer and linear equalizer, and derive closed-form expressions for the equalizer parameters and mean-square-error (MSE) for one-antenna systems. Then, we generalize the results to multiple-antenna systems through a *single channel equivalent model* [1]. Finally, we obtain the general configuration of the minimum MSE spatial-temporal equalizer for bandlimited systems, and show its application by a simulation example.**

Technical Area: Transmission and access systems

I. Introduction

Decision-feedback equalization [2], [3], [4] and linear equalization are effective techniques to remove intersymbol interference and co-channel interference. System performance can be further improved if antenna arrays are combined with the equalization. The structure and mean-square-error (MSE) of the optimum diversity combiner and decision-feedback equalizer (DFE) or linear equalizer (LE) have been derived in [1], [5], [6] for channels with additive white Gaussian noise. For channels with both additive Gaussian noise and co-channel interference, many researchers [7], [8], [9], [10] have investigated the optimum diversity combiner and DFE or LE from different points of view. In particular, for systems with one antenna, Peterson and Falconer [9], [10] have studied the minimum MSE (MMSE) DFE and LE for strictly bandlimited channels. In this paper, we analyze the performance of the MMSE spatial-temporal DFE (MMSE-STDFE) and LE (MMSE-STLE) for antenna array systems with co-channel interference, and derive closed-form expressions for the equalizer parameters and MMSE without this restriction. For bandlimited systems, as is always the case in wireless communications, we obtain a general configuration of MMSE spatial-temporal equalizer (STE).

II. System Model

For mobile wireless communication systems with M antennas, as shown in Figure 1, the received signal can be expressed in vector form as

$$\mathbf{x}(t)=\sum_{l=0}^{L}\sum_{n=-\infty}^{\infty}\mathbf{h}_l(t-nT)s_l[n]+\mathbf{n}(t),$$

with

$$\mathbf{x}(t)\triangleq[x_1(t),\cdots,x_M(t)]^T,$$
$$\mathbf{h}_l(t)\triangleq[h_l^{(1)}(t),\cdots,h_l^{(M)}(t)]^T,$$

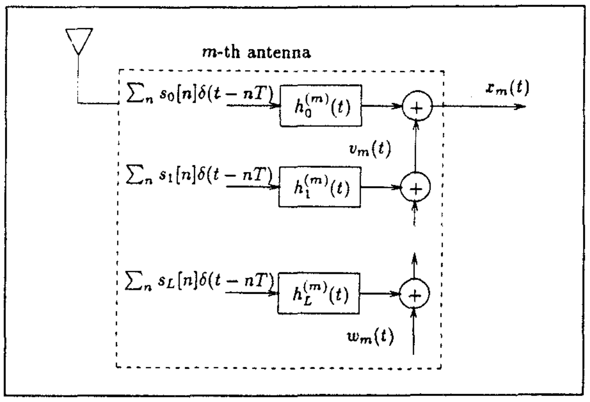


Fig. 1. System model.

and

$$\mathbf{n}(t)\triangleq[n_1(t),\cdots,n_M(t)]^T.$$

In the above expressions, T is the symbol period, $x_m(t)$ is the received signal from the m-th antenna, $h_0^{(m)}(t)$ is the combined channel and signal impulse response at the m-th antenna corresponding to the desired data and $h_l^{(m)}(t)$ is the combined impulse response of the m-th antenna corresponding to the l-th interferer, and $\{s_0[n]\}$ is the desired data from transmitter and $s_l[n]$, $l=1,\cdots,L$ is the complex data of the l-th interferer. We will assume that both the transmitted and the interference data are *independent, identically distributed* (i.i.d.) complex, zero-mean random variables with variance σ_s^2.

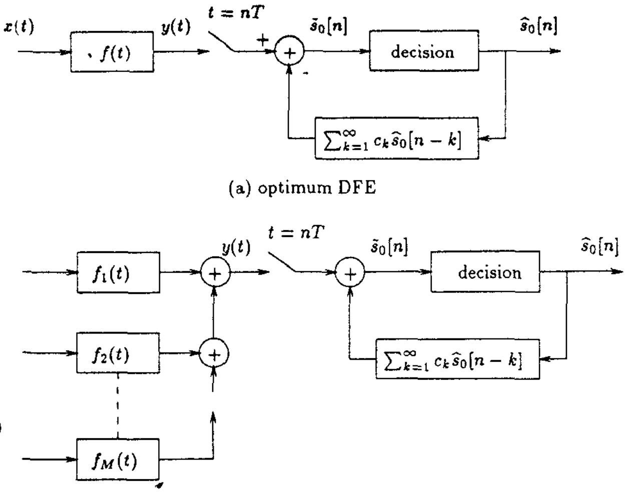
In some wireless communication systems, such as IS-136 TDMA systems, the shaping pulse $c(t)$ is a square-root raised-cosine with rolloff parameter β between 0 and 1. Therefore, the combined channel impulse response can be expressed as

$$h_l^{(m)}(t)=c(t)*g_l^{(m)}(t), \qquad (1)$$

where $*$ denotes convolution, and $g_l^{(m)}(t)$ represents the multipath fading of wireless channel.

III. Optimum DFE and LE for One-Antenna Systems with Co-channel Interference

Petersen and Falconer [9], [10] have investigated the structures and MSE's of the MMSE-DFE and MMSE-LE in the frequency domain for strictly band-limited channels. Below, we obtain closed-form expressions for the parameters and MSE's of the MMSE-DFE and MMSE-LE without this restriction.



(a) optimum DFE

(b) optimum spatial-temporal DFE

Fig. 2. MMSE-DFE for systems with cyclostationary interference.

The MMSE-DFE for a one-antenna system with cyclostationary interference is shown in Figure 2(a), which is similar to the MMSE-DFE for systems with stationary interference [4]. However, the expressions for $f(t)$ in the two cases are different although the derivation of the MMSE-DFE in both environments is similar. Here, we highlight the difference in the derivations.

Let the receiving filter $f(t)$ in Figure 2(a) have square-integrable impulse response $f(t)$. Then the output of the receiving filter is

$$y(t) \triangleq \int f(\tau)x(t-\tau)d\tau.$$

The output of the equalizer is

$$\begin{aligned}\tilde{s}[n] &\triangleq y(nT) - \sum_{k=1}^{\infty} c_k \hat{s}_0[n-k] \qquad (2)\\ &= \sum_{l=0}^{L}\sum_{k=-\infty}^{\infty} \int f(\tau)h_l(kT-\tau)d\tau s_l[n-k]\\ &\quad + \int f(\tau)w(nT-\tau)d\tau - \sum_{k=1}^{\infty} c_k \hat{s}_0[n-k].\end{aligned}$$

If the decided symbols are all correct, the intersymbol interference caused by $s_0[k]$ for $k < n$ can be eliminated by selecting

$$c_k = \int f(\tau)h_0(kT-\tau)d\tau.$$

If the data $s_l[n]$ are i.i.d. random variables, the MSE of the equalizer output is

$$\begin{aligned}e\{f(t)\} &\triangleq E\{|\tilde{s}[n]-s_0[n]|^2\}\\ &= \sigma^2 \left|\int f(\tau)h_0(-\tau)d\tau - 1\right|^2\\ &\quad +\sigma^2 \sum_{k=-\infty}^{-1} \left|\int f(\tau)h_0(kT-\tau)d\tau\right|^2\\ &\quad +\sigma^2 \sum_{l=1}^{L}\sum_{k=-\infty}^{\infty} \left|\int f(\tau)h_l(kT-\tau)d\tau\right|^2\\ &\quad +N_o \int |f(\tau)|^2. \qquad (3)\end{aligned}$$

Using calculus-of-variations, we can show that the $f(t)$ that minimizes the MSE satisfies

$$\begin{aligned}f_o(t) &= -\frac{1}{\tilde{N}_o}\{(a_0[0]-1)h_0^*(-t)\\ &\quad + \sum_{k=-\infty}^{-1} a_0[k]h_0^*(kT-t)\\ &\quad + \sum_{l=1}^{L}\sum_{k=-\infty}^{\infty} a_l[k]h_l^*(kT-t)\}, \qquad (4)\end{aligned}$$

where

$$a_l[k] \triangleq \int f_o(\tau)h_l(kT-\tau)d\tau, \quad \tilde{N}_o = \frac{N_o}{\sigma_s^2}. \qquad (5)$$

Multiplying both sides of (4) by $h_l(nT-\tau)$ and using (5), we have

$$\begin{aligned}a_i[n] &= -\{(a_0[0]-1)r_{i0}[n] + \sum_{k=-\infty}^{-1} a_0[k]r_{i0}[n-k]\\ &\quad + \sum_{j=1}^{L}\sum_{k=-\infty}^{\infty} a_j[k]r_{ij}[n-k]\}, \qquad (6)\end{aligned}$$

where

$$r_{ij}[n] \triangleq \frac{1}{\tilde{N}_o}\int h_i(nT+t)h_j^*(t)dt.$$

Let

$$r_{ij}(\omega) \triangleq \sum_{k=-\infty}^{\infty} r_{ij}[n]e^{-jk\omega}.$$

Using the Poisson sum formula[11], we have

$$r_{ij}(\omega) = \frac{1}{\tilde{N}_o T}\sum_{n=-\infty}^{\infty} H_i\left(\frac{\omega}{2\pi}-\frac{n}{T}\right)H_j^*\left(\frac{\omega}{2\pi}-\frac{n}{T}\right)$$

with

$$H_i(f) = \int_{-\infty}^{\infty} h_i(t)e^{-j2\pi ft}dt.$$

Denote the Fourier transform of the one-sided sequence $a_0[n] (n \leq 0)$ and the two-sided sequences $a_i[n]$, $i = 1, \cdots L$, respectively, as

$$a_0(\omega) \triangleq \sum_{k=-\infty}^{0} a_0[k]e^{-jk\omega}, \quad a_i(\omega) = \sum_{k=-\infty}^{\infty} a_i[k]e^{-jk\omega}.$$

Then (6), $i = 1, \cdots, L$ can be written in the frequency-domain as

$$a_i(\omega) = -\{(a_0(\omega)-1)r_{i0}(\omega) + \sum_{j=1}^{L} a_j(\omega)r_{ij}(\omega)\},$$

or in vector form as

$$\mathbf{a}(\omega) = -\{(a_0(\omega)-1)\mathbf{r}(\omega) + \mathbf{R}(\omega)\mathbf{a}(\omega)\}.$$

Therefore,

$$\mathbf{a}(\omega) = (1 - a_0(\omega))[\mathbf{R}(\omega)+\mathbf{I}]^{-1}\mathbf{r}(\omega), \qquad (7)$$

where $\mathbf{I}$ is an $L \times L$ identity matrix and

$$\mathbf{a}(\omega) = [a_1(\omega), \cdots, a_L(\omega)]^T,$$

$$\mathbf{r}(\omega) = [r_{10}(\omega), \cdots, r_{L0}(\omega)]^T,$$

$$\mathbf{R}(\omega) = (r_{ij}(\omega))_{i,j=1}^{3}.$$

Hence, $a_i[n]$, $i = 1, \cdots, L$ can be expressed in terms of $a_0[n]$ in the time-domain as

$$a_i[n] = b_i[n] - \sum_{k=-\infty}^{0} b_i[n-k]a_0[k], \qquad (8)$$

where the Fourier transform of $b_i[n]$ is the i-th element of the $L \times 1$ vector function

$$\mathbf{b}(\omega) = [\mathbf{R}(\omega)+\mathbf{I}]^{-1}\mathbf{r}(\omega).$$

When $i = 0$, (6) implies that,

$$\begin{aligned} a_0[n] &= -\{(a_0[0]-1)r_{00}[n] + \sum_{k=-\infty}^{-1} a_0[k]r_{00}[n-k] \\ &\quad + \sum_{j=1}^{L}\sum_{k=-\infty}^{\infty} a_i[k]r_{0i}[n-k]\}. \end{aligned} \qquad (9)$$

for $n \le 0$. By means of (8),

$$a_0[n] + \sum_{k=-\infty}^{0} r[n-k]a_0[k] = r[n],$$

where

$$r[n] = r_{00}[n] - v[n],$$

$$v[n] \triangleq \sum_{i=1}^{L} r_{0i}[n] * b_i[n] = \mathcal{F}^{-1}\{\mathbf{r}^H(\omega)[\mathbf{R}(\omega)+\mathbf{I}]^{-1}\mathbf{r}(\omega)\}.$$

Denote

$$r(\omega) \triangleq \sum_{n=-\infty}^{\infty} r[k]e^{-jn\omega} = r_{00}(\omega) - \mathbf{r}^H(\omega)[\mathbf{R}(\omega)+\mathbf{I}]^{-1}\mathbf{r}(\omega).$$

From Appendix A of [4],

$$a_0(\omega) = 1 - \frac{1}{M(-\omega)\gamma_0}, \qquad (10)$$

where $M(\omega)$ is a stable one-sided Fourier transform

$$M(\omega) = \sum_{n=0}^{\infty} \gamma_n e^{-jn\omega},$$

which is uniquely determined by

$$M(\omega)M(-\omega) = r(\omega) + 1.$$

The dc component γ_0 in $M(\omega)$ can be found by

$$\gamma_0^2 = \exp\{\frac{1}{2\pi}\int_{-\pi}^{\pi} \ln[(r(\omega)+1]d\omega\}.$$

Substituting (10) into (7), we have

$$\mathbf{a}(\omega) = \frac{1}{M(-\omega)\gamma_0}[\mathbf{R}(\omega)+\mathbf{I}]^{-1}\mathbf{r}(\omega).$$

Multiplying both sides of (4) by $f_o(t)$ and integrating, from (3), the MSE of the MMSE-DFE is

$$\begin{aligned} e\{f_o(t)\} &= \sigma^2(1 - a_0[0]) \\ &= \sigma^2 \exp\{-\frac{1}{2\pi}\int_{-\pi}^{\pi} \ln\{r(\omega)+1\}d\omega\}. \end{aligned} \qquad (11)$$

Following a similar derivation, the $f(t)$ and MSE for the MMSE-LE are

$$f_o(t) = \frac{1}{\tilde{N}_0}\{h_0^*(-t) - \sum_{l=0}^{L}\sum_{k=-\infty}^{\infty} a_l[k]h_l^*(kT-t)\}, \qquad (12)$$

and

$$e\{f_o(t)\} = \frac{\sigma_s^2}{2\pi}\int_{-\pi}^{\pi} \frac{1}{r(\omega)+1}d\omega. \qquad (13)$$

The parameter $a_l[n]$ for the MMSE-LE is given by

$$a_0(\omega) = \frac{r(\omega)}{1+r(\omega)}, \ \mathbf{a}(\omega) = \frac{r(\omega)}{1+r(\omega)}[\mathbf{R}(\omega)+I]^{-1}\mathbf{r}(\omega).$$

where the definitions of $a_l(\omega)$, $l = 1, \cdots, L$, and $\mathbf{a}(\omega)$ are the same as before, except that $a_0(\omega)$ is the two-sided Fourier transform defined as

$$a_0(\omega) \triangleq \sum_{n=-\infty}^{\infty} a_0[n]e^{-jn\omega}.$$

IV. Optimum STDFE and STLE for Multiple Antenna Systems with Co-channel Interference

Using the *single channel equivalent model* developed in [1], we can easily extend the above results to multiple-antenna systems to derive the MMSE-STDFE and MMSE-STLE with cyclostationary interference.

For an M-antenna system, we first define the *compounded channel impulse responses* $h_l(t)$ and the *compounded channel additive noise* $n(t)$, respectively, as

$$h_l(t) = \sqrt{M}h_l^{(m)}(nT + M(t - nT - \frac{m-1}{M}T)), \qquad (14)$$

$$n(t) = \sqrt{M}n_m(nT + M(t - nT - \frac{m-1}{M}T)), \qquad (15)$$

for $nT + \frac{m-1}{M}T \le t < nT + \frac{m}{M}T$, $l = 0, 1, \cdots, L$, $m = 1, \cdots, M$, and $n = 0, \pm1, \pm2, \cdots$. According to

[1], a single-antenna system with desired signal channel impulse response $h_0(t)$, interference channel impulse responses $h_l(t)$, $l = 1, \cdots, L$, and additive noise $n(t)$ is equivalent to the M-antenna system.

From the results established in the previous section, the $f(t)$ for the MMSE-DFE and MMSE-LE can be expressed as (4) and (12) respectively. Let

$$\mathbf{f}(t) \triangleq [f_1(t), \cdots, f_m(t)]^T.$$

Hence, by virtue of (14), the $\mathbf{f}(t)$ for the MMSE-STDFE in Figure 2(b) is

$$\begin{aligned} \mathbf{f}_o(t) &= \frac{1}{\tilde{N}_o}\{\mathbf{h}_0^*(-t) - \sum_{k=-\infty}^{0} a_0[k]\mathbf{h}_0^*(kT-t) \\ &\quad - \sum_{l=1}^{L}\sum_{k=-\infty}^{\infty} a_l[k]\mathbf{h}_l^*(kT-t)\}. \end{aligned} \tag{16}$$

The $\mathbf{f}(t)$ for the MMSE-STLE is

$$\mathbf{f}_o(t) = \frac{1}{\tilde{N}_o}\{\mathbf{h}_0^*(-t) - \sum_{l=0}^{L}\sum_{k=-\infty}^{\infty} a_l[k]\mathbf{h}_l^*(kT-t)\}.$$

The expressions of the parameter $a_i[n]$ and MSE for the MMSE-STDFE and MMSE-STLE are the same as those in the previous section except that $r_{ij}(\omega)$ is replaced by

$$r_{ij}(\omega) \triangleq \frac{1}{\tilde{N}_o T}\sum_{m=1}^{M}\sum_{n=-\infty}^{\infty} H_i^{(m)}(\frac{\omega}{2\pi} - \frac{n}{T})H_j^{(m)*}(\frac{\omega}{2\pi} - \frac{n}{T}),$$

where

$$H_i^{(m)}(f) = \int_{-\infty}^{\infty} h_i^{(m)}(t)e^{-j2\pi f t}dt.$$

Since $\mathbf{h}_l(t)$, $l = 0, 1, \cdots, L$, usually differ, the concept of the *matched filter* for stationary interference systems is not valid here.

Note that, if there is no cyclostationary interference, then the $\mathbf{f}_o(t)$ and the minimum MSE are the same as those in [5].

V. General Configuration of the MMSE-STE for Bandlimited Systems

Let $s(t)$ be any $1/T$ band limited signal whose spectrum satisfies

$$S(f) = \begin{cases} 1 & |f| \leq (1+\beta)/2T \\ 0 & |f| \geq 1/T \end{cases}, \tag{17}$$

Note that $S(f)$, $(1+\beta)/2T < |f| < 1/T$, can take any value to make it square integrable. Since $s(t)$ is $1/T$-band-limited, $c(t) * s(t) = c(t)$. Hence, for $l = 1, \cdots, L$ and $m = 1, \cdots, M$,

$$h_l^{(m)}(t) = c(t) * g_l^{(m)}(t) = c(t) * s(t) * g_l^{(m)}(t).$$

Using the *Sampling Theorem*, we have

$$\begin{aligned} s(t) * g_l^{(m)}(t) &= \sum_{n=-\infty}^{\infty} g_l^{(m)}[n]s_o(t - n\frac{T}{2}) \\ &= s_o(t) * \{\sum_{n=-\infty}^{\infty} g_l^{(m)}[n]\delta(t - n\frac{T}{2})\}, \end{aligned} \tag{18}$$

where

$$g_l^{(m)}[n] \triangleq \frac{T}{2}\int_{-\infty}^{\infty} s(n\frac{T}{2} - \tau)g_l^{(m)}(\tau)d\tau,$$

$$s_o(t) \triangleq \frac{2}{T}\frac{\sin(2\pi t/T)}{2\pi t/T}, \text{ and } \mathcal{F}\{s_o(t)\} = \begin{cases} 1 & |f| \leq 1/T \\ 0 & |f| > 1/T \end{cases}.$$

Hence,

$$\begin{aligned} h_l^{(m)}(t) &= c(t) * s_o(t) * \{\sum_{n=-\infty}^{\infty} g_l^{(m)}[n]\delta(t - n\frac{T}{2})\} \qquad (19) \\ &= c(t) * \{\sum_{n=-\infty}^{\infty} g_l^{(m)}[n]\delta(t - n\frac{T}{2})\}, \\ &= c(t) * \{\sum_{n=-\infty}^{\infty} g_l^{(m)}[2n]\delta(t - nT)\} \\ &\quad + c(t - \frac{T}{2}) * \{\sum_{n=-\infty}^{\infty} g_l^{(m)}[2n+1]\delta(t - nT)\} \end{aligned}$$

Substituting the above identity into (16), we have

$$\begin{aligned} f_o^{(m)}(t) &= c(t) * \{\sum_{n=-\infty}^{\infty} f_{1,m}[n]\delta(t - nT)\} \\ &\quad + c(t - \frac{T}{2}) * \{\sum_{n=-\infty}^{\infty} f_{2,m}[n]\delta(t - nT)\}, \end{aligned} \tag{20}$$

where

$$\begin{aligned} f_{m,1}[n] &= \frac{1}{\tilde{N}_o}\{-g_l^{(m)*}[-2n] \\ &\quad + \sum_{k=-\infty}^{0} a_0[k]g_0^{(m)*}[2k-2n] \\ &\quad + \sum_{l=1}^{L}\sum_{k=-\infty}^{\infty} a_l[k]g_l^{(m)*}[2k-2n]\}, \end{aligned}$$

and

$$\begin{aligned} f_{m,2}[n] &= \frac{1}{\tilde{N}_o}\{-g_l^{(m)*}[-2n+1] \\ &\quad + \sum_{k=-\infty}^{0} a_0[k]g_0^{(m)*}[2k-2n+1] \\ &\quad + \sum_{l=1}^{L}\sum_{k=-\infty}^{\infty} a_l[k]g_l^{(m)*}[2k-2n+1]\}. \end{aligned}$$

Hence, the MMSE-STDFE in Figure 2(b) can be implemented as in Figure 3, where $F_{m,i}$ are discrete filters with parameters $f_{m,i}[n]$.

It can be shown that the MMSE-STLE has a similar structure to that in Figure 3, but without the decision-feedback filter.

VI. A Simulation Example of STE

The performance of the STE has been evaluated through computer simulation, which focused on its application in IS-136 TDMA systems. The simulation uses the system model

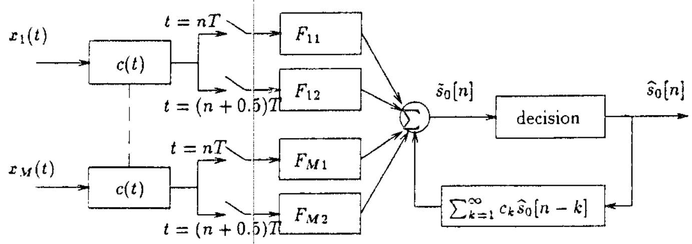


Fig. 3. General configuration of MMSE-STDFE for systems with cyclostationary interference.

described in Section II. Each time slot contains a 14 symbol training sequence followed by 134 symbols randomly drawn from $\{\frac{1+j}{\sqrt{2}}, \frac{1-j}{\sqrt{2}}, \frac{-1+j}{\sqrt{2}}, \frac{-1-j}{\sqrt{2}}\}$. The parameters of the equalizers are initially estimated using the training sequence, and after the training period, they are tracked using decided (sliced) symbols. DQPSK modulation is used with coherent detection. The 4-antenna system has white Gaussian noise and a single co-channel interferer, whose powers are given by the signal-to-noise ratio (SNR) and the signal-to-interference ratio (SIR), respectively. The channels use the two-path model with the same average power for each path, the same delay spread for both desired and interference channels, and f_d = 184 Hz. The signal received by each antenna is first passed through a square-root raised-cosine filter and then oversampled at the ideal sampling time at a rate of $2/T$ for the STE. The desired signal and interference are time-aligned for the results presented in this section (Note that the relative timing does not significantly affect the performance of the STE). One feedback tap is used for the STDFE. To give insight into the average behavior of the STE in various environments, we have averaged the performance over 1,000 time slots.

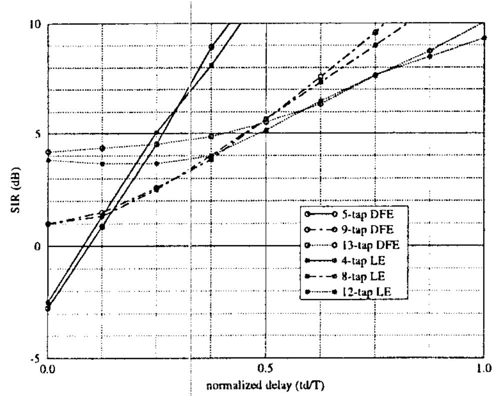


Fig. 4. Performance of DLMMSE-STDFE and DLMMSE-STLE: required SIR for BER= 10^{-2} versus t_d with f_d = 184 Hz and SNR= 20 dB.

Figure 4 shows the required SIR for a BER=10^{-2} of different length DLMMSE-STE's for channels with SNR= 20 dB and different t_d's. From the figure, without delay spread, both the 5-tap DLMMSE-DFE and 4-tap DLMMSE-LE, *i.e.*, spatial processing only, operate up to -2.5dB SIR. With increasing t_d, the equalizer's interference suppression ability is reduced. As t_d increases, the equalizer performance is generally improved by increasing the number of taps. However, for rapid dispersive fading channels, a too long equalizer does not necessarily have good performance because the parameter tracking performance degrades with increasing equalizer length, even through the longer equalizer always performs better than the shorter one with the optimum equalizer parameters. Hence, in Figure 4, the 5-tap DLMMSE-DFE and 4-tap DLMMSE-LE have the best performance if $t_d \leq T/8$, while the 13-tap DLMMSE-STDFE and 12-tap DLMMSE-STLE have the best performance if $t_d > T/2$. Usually $t_d < T/2$ in IS-136 TDMA systems [12], therefore, the 9-tap DLMMSE-STDFE and 8-tap DLMMSE-STLE are two of the best STE's

VII. Conclusions

In this paper, we derived the structures and the MSE of optimum diversity combiner and decision-feedback/linear equalizers for diversity receiving systems with both additive stationary noise and cyclostationary interference. For bandlimited systems, we obtained a general configuration for the MMSE STE. As shown by a simulation example, the MMSE STE can be used in wireless mobile systems to mitigate intersymbol interference and suppress co-channel interference.

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