PAPER Analog Single-Carrier Transmission with Frequency-Domain Equalization

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SUMMARY In this paper, a new analog signal transmission technique called analog single-carrier transmission with frequency-domain equalization (analog SC-FDE) is proposed. Analog SC-FDE applies discrete Fourier transform (DFT), frequency-domain spectrum shaping and mapping, inverse DFT (IDFT), and cyclic prefix (CP) insertion before transmission. At the receiver, one-tap FDE is applied to take advantage of frequency diversity. This paper considers, as an example, analog voice transmission. A theoretical analysis of the normalized mean square error (NMSE) performance is carried out to evaluate the transmission property of the proposed analog SC-FDE and is confirmed by computer simulation. We show that analog SC-FDE achieves better NMSE performance than conventional analog signal transmission scheme.

key words: analog signal transmission, frequency-domain equalization, single-carrier transmission

1. Introduction

Nowadays, although digital signal transmission has been continuously evolving [1]–[3], analog signal transmission (e.g., radio broadcasting) still remains essential. In comparison with the digital signal transmission, a narrower occupied bandwidth is achieved in the analog signal transmission since the analog signal transmission does not need any source coding nor channel coding. For example, to transmit 4 kHz bandwidth of analog voice signal, digital transmission system, which does not include channel coding, needs 64 kbps for pulse code modulation (PCM) [4] in case of using 8-bit source coding. Even a high-efficiency speech coding has been recently applied in GSM or W-CDMA networks, the systems also need a broader bandwidth with source coding rate (not including channel coding) of $4.75 \sim 12.2$ kbps [5].

Because of the narrowband transmission, the channel in analog signal transmission is considered as frequencynonselective fading channel [3]. As a consequence of suffering from the frequency-nonselective fading channel, the received signal power drops over a consecutive period of time and hence, the received signal quality significantly degrades. In order to overcome this problem, we focus on a solution that widens the bandwidth of the analog signal by using discrete Fourier transform (DFT) to utilize the advantage of frequency-selective fading channel (i.e., frequency

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diversity). The widening of signal bandwidth can be done by applying DFT and mapping the frequency components over a broader bandwidth. However, in such the frequencyselective fading channel, the received signal spectrum is severely distorted and thus, transmission performance degrades. This means that some techniques that can correct the spectrum distortion need to be adopted.

For broadband digital signal transmission, it is well known that orthogonal frequency division multiplexing (OFDM) [6] is robust against the frequency-selective fading channel, but its high peak-to-average power ratio (PAPR) of the transmitted signal is the main drawback [7]. By comparison, single-carrier (SC) transmission has lower PAPR while achieving good transmission performance by using frequency-domain equalization (FDE) at the receiver [8]– [14]. As a consequence, SC-FDE has recently drawn great attention as a robust digital signal transmission technique.

In this paper, we apply SC-FDE technique in order to improve the performance of analog signal transmission and propose a new analog signal transmission technique called analog SC transmission with FDE (analog SC-FDE). The proposed analog SC-FDE applies DFT, frequency-domain spectrum shaping and mapping, inverse DFT (IDFT), and cyclic prefix (CP) insertion before transmission. At the receiver, one-tap FDE is applied to take advantage of the channel frequency-selectivity. A theoretical analysis of the normalized mean square error (NMSE) performance is carried out to evaluate the proposed analog SC-FDE and is confirmed by computer simulation. We show that the proposed analog SC-FDE achieves better NMSE performance than conventional analog signal transmission. In this paper, single channel transmission is considered. However, multiaccess is also possible based on the principle of SC frequencydivision multiple access (SC-FDMA) [15]. Therefore, the spectrum efficiency is the same as conventional analog signal transmission, inherently narrowband.

Although PCM or high-efficiency speech coding in the digital signal transmission has a delay for source coding, analog SC-FDE does not have this delay. The time delay for DFT/IDFT processing in analog SC-FDE is the same as that of digital SC-FDE transmission. However, in comparison with conventional analog signal transmission, analog SC-FDE has a particular time delay due to the block signal processing. For example, by considering a sampling rate of 8 kHz and a block size of 64 samples (as assumed in computer simulation in this paper), this time delay is approxi-

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mately 8 ms.

The remainder of this paper is organized as follows. In Sect. 2, we propose the system model of the analog SC-FDE and describe the principle. In Sect. 3, an expression for the NMSE performance of analog SC-FDE is derived with a given set of channel gains. Next, Sect. 4 shows computer simulation and theoretical results to confirm the effectiveness of the proposed analog SC-FDE. Finally, Sect. 5 provides the conclusion and future work.

2. Analog SC-FDE

2.1 Transmission System Model

The transmission system model of analog SC-FDE is illustrated in Fig. 1. At the transmitter, after the signal bandwidth is limited by low-pass filter (LPF), the analog signal s(t) to be transmitted is sampled at the Nyquist rate. Then, the sample sequence is grouped into a sequence of signal blocks of M samples each. Each signal block $\{s(n); n=0 \sim M-1\}$ is transformed by M-point DFT into frequency-domain signal block. Spectrum shaping filter is introduced in order to generalize the proposed analog SC-FDE having a specific spectrum shaping design. For example, because of the complex conjugate relationship between frequency components in the upper sideband (USB) and lower sideband (LSB), spectrum shaping filter can be designed to remove one of these sidebands and transmit only a half of analog signal spectrum (i.e., a novel scheme of single-sideband (SSB) transmission without Hilbert transform). The receiver can restore the original spectrum by using a filter which is correspondent with spectrum shaping filter. Therefore, spectrum efficiency of analog SC-FDE transmission increases twofold. Additionally, spectrum shaping filter can be also designed for a low-frequency or high-frequency emphasis if necessary. In this paper, we assume an ideal brick wall LPF as the spectrum shaping filter.



Fig. 1 Transmission system model of analog SC-FDE.

The resultant M frequency components are mapped over a broad bandwidth having $N_c(>M)$ orthogonal subcarriers with zeros occupying the unused subcarriers. Then, the resultant frequency domain signal of N_c subcarriers is transformed by N_c -point IDFT back into complex time-domain signal block { $x(n); n = 0 \sim N_c - 1$ }. Finally, the last N_g samples of the complex time-domain signal block are copied as a CP and inserted into the guard interval (GI) placed at the

beginning of each transmit signal block. The CP-inserted signal block is transmitted over a frequency-selective fading channel. At the receiver, CP is removed from the each received signal block and then, each block is transformed by N_c -point DFT into N_c subcarrier components. After performing de-mapping and FDE, *M* subcarriers of original signal are picked up and transformed by *M*-point IDFT back into complex signal block $\{\tilde{s}(n); n = 0 \sim M - 1\}$ of *M* samples. Finally, the analog signal $\tilde{s}(t)$ is reconstructed after applying the automatic gain control (AGC) [16] and LPF.

2.2 Subcarrier Mapping

At the transmitter, we assume that spectrum shaping filter is an ideal brick wall LPF. *M* frequency components after applying spectrum shaping filter are denoted by $\{S(k); k = 0 \sim M - 1\}$ and expressed as

$$S(k) = \frac{1}{\sqrt{M}} \sum_{n=0}^{M-1} s(n) \exp\left(-j2\pi k \frac{n}{M}\right).$$
 (1)

The transmitter then maps M frequency components over a broad bandwidth having $N_c(> M)$ orthogonal subcarriers expressed as $\{X(k); k = 0 \sim N_c - 1\}$. We consider two mapping modes: localized mapping and distributed mapping [13] shown in Fig. 2 in the case of M = 6 and $N_c = 24$.

• Localized mapping

$$X(k) = \begin{cases} S(k), & k = 0 \sim M - 1\\ 0, & \text{otherwise} \end{cases}$$
(2)





Fig.2 Subcarrier mapping in analog SC-FDE and conventional analog FDMA.

Distributed mapping

$$X(k) = \begin{cases} S(k'), & k = k' \times \frac{N_c}{M} \\ 0, & \text{otherwise} \end{cases}$$
(3)

In Eqs. (2) and (3), $k'=0 \sim M-1$, $k=0 \sim N_c-1$, and N_c/M is the adjacent subcarrier interval. In localized mode, M frequency components are mapped to M contiguous subcarriers. On the other hand, in distributed mode, they are mapped to equally-spaced M subcarriers. In both modes, zeros occupy the unused subcarriers. In case of transmitting multiple analog signal streams, each consisting of M subcarriers, the subcarrier mapping is performed so that multiple streams do not overlap (or they are orthogonal) one another in the frequency-domain similar to the principle of SC-FDMA. As an application, analog SC-FDE can be applied to existing analog systems such as radio broadcasting. In this case, all channels transmit the same whole broad bandwidth which is allocated to the radio broadcasting (e.g., bandwidth of approximately 1 MHz for AM radio system in Japan). This is the different point to existing radio broadcasting in which each channel only transmits the narrowband signal shown in Fig. 2(c). For example, in multi-channels (N channels) radio broadcasting system using analog SC-FDE, each channel maps M frequency components of original signal over the same broader bandwidth having $N_c(=M \times N)$ orthogonal subcarriers as the other channels. The subcarrier mapping of all channels is performed so that frequency components of each original signal do not overlap one another in the frequency-domain, as shown in Figs. 2(a) and (b). Conventional analog FDMA is also illustrated in Fig. 2 to show that analog SC-FDE has the same spectrum efficiency as conventional analog FDMA.

After N_c -point IDFT, the time-domain sample sequence at a rate of $1/T_s = (N_c/M) \times 1/T$, in which 1/T is the Nyquist sampling rate of analog signal s(t), is obtained. The CP-inserted time-domain sample sequence $\{\tilde{x}(n); n = -N_g \sim N_c - 1\}$ can be expressed using the equivalent low-pass representation as

$$\tilde{x}(n) = \sqrt{2P} x \left(n \mod N_c \right), \tag{4}$$

where *P* is the average sample sequence power and $\{x(n); n = 0 \sim N_c - 1\}$ is given by

$$x(n) = \frac{1}{\sqrt{N_c}} \sum_{k=0}^{N_c - 1} X(k) \exp\left(j2\pi n \frac{k}{N_c}\right).$$
 (5)

2.3 Received Signal

Assuming that the channel consists of *L* distinct propagation paths, the channel impulse response $h(\tau)$ can be expressed as

$$h(\tau) = \sum_{l=0}^{L-1} h_l \,\delta(\tau - \tau_l)\,, \tag{6}$$

where h_l and τ_l are the complex-valued path gain with

 $E\left[\sum_{l=0}^{L-1} |h_l|^2\right] = 1$ (*E*[.] denotes the expectation operation) and the *l*-th path delay time, respectively. In Eq. (6), we assume that the channel stays constant during the signal transmission period of one block. It is assumed that the maximum time delay of channel is shorter than CP and the received signal is ideally sampled at the rate $1/T_s$. The discrete-time received signal {r(t); $t = -N_g \sim N_c - 1$ } is

$$r(t) = \sum_{l=0}^{L-1} h_l \,\tilde{x} \,(t - \tau_l) + n(t), \tag{7}$$

where n(t) is the additive white Gaussian noise (AWGN) with zero-mean and variance $2N_0/T_s$ in which N_0 is the single-sided power spectrum density.

After removing CP, the receiver transforms the received signal block into the frequency-domain signal using N_c -point DFT. The frequency-domain received signal $\{R(k); k=0 \sim N_c-1\}$ can be expressed as

$$R(k) = \sqrt{2P} H(k)X(k) + \Pi(k), \qquad (8)$$

where H(k) and $\Pi(k)$ are the channel gain and the noise component at the *k*-th frequency, respectively. They are given by

$$H(k) = \sum_{l=0}^{L-1} h_l \exp\left(-j2\pi k \frac{\tau_l}{N_c}\right).$$
(9)

$$\Pi(k) = \frac{1}{\sqrt{N_c}} \sum_{t=0}^{N_c-1} n(t) \exp\left(-j2\pi k \frac{t}{N_c}\right).$$
 (10)

2.4 Subcarrier De-Mapping and FDE

De-mapping is performed to obtain desired *M* frequency components $\{\hat{R}(k); k=0 \sim M-1\}$ of original signal. Channel gain $\{\hat{H}(k); k=0 \sim M-1\}$ for FDE and the equivalent noise component $\{\hat{\Pi}(k); k=0 \sim M-1\}$ are also obtained as follows. In this paper, ideal channel estimation is assumed.

· Localized de-mapping

$$\begin{cases} \hat{R}(k) = R(k) \\ \hat{H}(k) = H(k), \ k = 0 \sim M - 1. \\ \hat{\Pi}(k) = \Pi(k) \end{cases}$$
(11)

• Distributed de-mapping

$$\begin{cases} \hat{R}(k) = R(k \times N_c/M) \\ \hat{H}(k) = H(k \times N_c/M), \ k = 0 \sim M - 1. \\ \hat{\Pi}(k) = \Pi(k \times N_c/M) \end{cases}$$
(12)

After the subcarrier de-mapping, one-tap FDE is carried out as

$$\widetilde{S}(k) = W(k)\widehat{R}(k) = \sqrt{2PW(k)\widehat{H}(k)S(k)} + W(k)\widehat{\Pi}(k), \quad (13)$$

where W(k) is the FDE weight. We consider three FDE weights based on zero-forcing (ZF) criterion, maximal-ratio combining (MRC) criterion and minimum mean square error (MMSE) criterion [14] given as

$$W(k) = \begin{cases} \frac{1}{\hat{H}(k)}, & ZF \\ \hat{H}^{*}(k), & MRC \\ \frac{\hat{H}^{*}(k)}{\left|\hat{H}(k)\right|^{2} + \Gamma^{-1}}, & MMSE \end{cases}$$
(14)

where $\Gamma = PT_s/N_0$ is the average signal-to-noise power ratio (SNR) and [.]* denotes the complex conjugate operation.

After transforming the frequency-domain signal back into the time-domain signal by *M*-point IDFT, only the real part of the time-domain signal $\{\tilde{s}(n); n=0 \sim M-1\}$ is outputted as

$$\tilde{s}(n) = \frac{1}{K} \times \operatorname{Re}\left\{\frac{1}{\sqrt{M}} \sum_{k=0}^{M-1} \tilde{S}(k) \exp\left(j2\pi n \frac{k}{M}\right)\right\}$$
$$= s(n) + \operatorname{Re}\left\{\mu_{\mathrm{ISI}}(n) + \mu_{\mathrm{noise}}(n)\right\}, \qquad (15)$$

where *K*, $\mu_{ISI}(n)$, and $\mu_{noise}(n)$ are the normalization factor of AGC, residual signal distortion, and equivalent noise, respectively, which are given by

$$\begin{cases} K = \sqrt{2P} \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \\ \mu_{\rm ISI}(n) = \frac{1}{K} \left\{ \sqrt{2P} \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \left[\sum_{\substack{n'=0\\n'\neq n}}^{M-1} s(n') \exp\left(j2\pi k \frac{n-n'}{M}\right) \right] \right\}, \\ \mu_{\rm noise}(n) = \frac{1}{K} \left\{ \frac{1}{\sqrt{M}} \sum_{k=0}^{M-1} \tilde{\Pi}(k) \exp\left(j2\pi n \frac{k}{M}\right) \right\}$$
(16)

with $\tilde{H}(k) = W(k)\hat{H}(k)$ and $\tilde{\Pi}(k) = W(k)\hat{\Pi}(k)$ being the equivalent channel and the equivalent noise component at the *k*-th frequency, respectively. Finally, the analog signal $\tilde{s}(t)$ is reconstructed by LPF from discrete-time signal $\tilde{s}(n)$.

2.5 Conventional Analog Signal Transmission

For comparison, conventional double sideband (DSB) [17] transmission is considered. At the transmitter, after applying LPF, the analog signal is transmitted without any signal processing. Due to the narrow bandwidth of signal, we assume that the channel consists of one path (i.e., frequency-nonselective fading channel).

At the receiver, ideal fast AGC suppressing the fluctuation of received signal power due to fading and ideal coherent demodulation are assumed. The continuous-time representation of the DSB demodulated signal can be expressed as

$$\tilde{s}(t) = s(t) + \operatorname{Re}\left\{\eta(t) \left| \left(\sqrt{2P} h(t)\right) \right\},\tag{17}$$

where h(t), s(t), and $\eta(t)$ are complex-valued path gain, transmitted signal, and the zero-mean complex-valued AWGN having the double-sided power spectrum density $2N_0$, respectively.

3. Normalized Mean Square Error (NMSE) Analysis

The demodulated signal $\{\tilde{s}(n); n = 0 \sim M - 1\}$ at the analog SC-FDE receiver is expressed by Eq. (15). In this paper, in order to evaluate transmission performance of the proposed analog SC-FDE, we use NMSE criterion which is defined as

NMSE =
$$\frac{E\left[\left|\tilde{s}(n) - s(n)\right|^{2}\right]}{E\left[\left|s(n)\right|^{2}\right]}.$$
(18)

Without loss of generality, the transmit signal is assumed to have unit average power. Using Eq. (15), NMSE can be written as

$$\text{NMSE} = E\left[|\text{Re}\left\{\mu_{\text{ISI}}(n) + \mu_{\text{noise}}(n)\right\}|^2\right].$$
 (19)

Since $\mu_{ISI}(n)$ and $\mu_{noise}(n)$ are statistically independent, the variance of $\mu(n) = \mu_{ISI}(n) + \mu_{noise}(n)$ is expressed as

$$2\sigma^2 = E\left[|\mu(n)|^2\right] = 2\sigma_{\rm ISI}^2 + 2\sigma_{\rm noise}^2,\tag{20}$$

where $\sigma_{\rm ISI}^2$ and $\sigma_{\rm noise}^2$ are derived as (see Appendix)

$$\sigma_{\rm ISI}^{2} = \frac{1}{2} \left\{ \frac{\frac{1}{M} \sum_{k=0}^{M-1} \left| \tilde{H}(k) \right|^{2}}{\left| \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \right|^{2}} - 1 \right\},$$
(21)
$$\sigma_{\rm noise}^{2} = \frac{1}{2} \cdot \frac{\Gamma^{-1} \cdot \frac{1}{M} \sum_{k=0}^{M-1} |W(k)|^{2}}{\left| \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \right|^{2}}.$$
(22)

Therefore, the conditional NMSE for the given set of $\{H(k); k = 0 \sim N_c - 1\}$ (or equivalently, the given set of path gains $\{h_l; l = 0 \sim L - 1\}$) can be expressed as

NMSE
$$(\Gamma, \{H(k)\}) = \sigma^2 = \sigma_{ISI}^2 + \sigma_{noise}^2$$

= $\frac{1}{2} \left\{ \frac{\frac{1}{M} \sum_{k=0}^{M-1} |\tilde{H}(k)|^2 + \Gamma^{-1} \cdot \frac{1}{M} \sum_{k=0}^{M-1} |W(k)|^2}{\left| \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \right|^2} - 1 \right\}.$ (23)

The theoretical average NMSE can be numerically evaluated by averaging Eq. (23) over $\{H(k); k=0 \sim N_c - 1\}$ as

NMSE (
$$\Gamma$$
) = \int NMSE (Γ , { $H(k)$ }) $\prod_{k} dH(k)$. (24)

4. Simulation and Theoretical Results

Analog SC-FDE takes an advantage of channel frequencyselectivity to improve the transmission performance. Therefore, as the maximum delay time difference or propagation path length difference gets longer, the performance improves. In this paper, as an example, a propagation channel

Signal transmission	Analog SC-FDE		Conventional DSB
Total no. of subcarriers	$N_c = 8192$		
Time-domain block size	M = 64		
No. of channels	N = 1		
Spectrum shaping filter	Ideal brick wall LPF		
GI length	$N_{g} = 16$		
Sampling rate	1/T = 8 k		хHz
Mapping	Localized	Distributed	
Channel	Frequency-selective block Rayleigh fading L=16-path uniform power delay profile $\tau_l = l$		Frequency non- selective Rayleigh fading
FDE weight	ZF, MRC, MMSE		
Channel estimation & Fast AGC	Ideal		

 Table 1
 Computer simulation condition.

model of 16 paths is assumed with the maximum delay time difference of approximately $15\,\mu$ s (which is equivalent to a maximum propagation path length difference of 4.5 km). Analog SC-FDE can work in other propagation conditions having longer delay time difference than $15\,\mu$ s. However, when the distributed mapping is used, the performance difference is not so sensitive to the maximum delay time. It should be noted that longer GI length is necessary.

4.1 Computer Simulation Condition

The condition for numerical evaluation of the theoretical average NMSE and the computer simulation is summarized in Table 1. In the proposed analog SC-FDE, we assume the bandwidth-limited (4 kHz) voice transmission. A sampling rate of 8 kHz, a time-domain signal block with length of M = 64 samples, an adjacent subcarrier interval of 125 Hz, and a subcarrier mapping (localized and distributed) over $N_c = 8192$ subcarriers are assumed. As a propagation channel, a frequency-selective block Rayleigh fading channel having an L=16-path uniform power delay profile is considered. The *l*-th path time delay is $\tau_l = l$ and the maximum delay difference is less than GI length (i.e., $L-1 \le N_a$). Ideal channel estimation is assumed. On the other hand, in conventional analog signal transmission (i.e., conventional DSB), the channel is assumed as a frequency-nonselective fading channel and ideal fast AGC is also assumed.

The numerical evaluation of the theoretical average NMSE is done by Monte-Carlo numerical computation method as follows. The set of path gains $\{h_l; l = 0 \sim L-1\}$ is generated for obtaining $\{H(k); k = 0 \sim N_c - 1\}$ and $\{W(k); k = 0 \sim M - 1\}$ using Eqs. (9) and (14), respectively. The conditional NMSE for the given average SNR Γ is computed using Eq. (23). This is repeated sufficient times to obtain the average NMSE given by Eq. (24).

4.2 NMSE Performance

The NMSE performance of the proposed analog SC-FDE in the case of cosine wave transmission and voice transmission



Fig.3 NMSE performance of analog SC-FDE in the case of cosine wave transmission.



Fig. 4 NMSE performance of analog SC-FDE in the case of voice transmission.

are plotted in Fig. 3 and Fig. 4, respectively. For comparison, NMSE performance of the conventional DSB transmission is also plotted. It can be seen that both of cosine wave and voice transmission have almost the same performance. The proposed analog SC-FDE achieves much better performance than conventional DSB. Among three FDE weights, SC-FDE using ZF-FDE provides almost the same NMSE performance as the conventional DSB. MMSE-FDE with distributed mapping achieves the best performance for all average SNRs. The reasons are discussed below.

Using the ZF weight (i.e., $W(k) = \hat{H}^{-1}(k)$), the equivalent channel becomes flat (i.e., $\tilde{H}(k) = 1$, $k = 0 \sim M - 1$) and as a consequence, the residual signal distortion disappears

in Eq. (15) (i.e., $\mu_{ISI}(n) = 0$). Therefore, the comparison of $\tilde{s}(n)$ in Eq. (15) and $\tilde{s}(t)$ in Eq. (17) shows that the proposed analog SC-FDE provides the same output as the conventional DSB. Consequently, the proposed analog SC-FDE using ZF-FDE has almost same performance as conventional DSB. However, it should be noted that the noise of analog SC-FDE using ZF-FDE and the noise of conventional DSB using fast AGC are both enhanced when the channel gain $\hat{H}(k)$ or h(t) drops.

The MRC weight (i.e., $W(k)=\hat{H}^*(k)$) can avoid the noise enhancement problem and maximizes the average SNR, but it enhances the frequency-selectivity of the equivalent channel (the channel after FDE) in the case of distributed mapping and thus, the distributed mapping with MRC-FDE cannot improve the performance even when the average SNR increases. But in the case of localized mapping, MRC-FDE does not really enhance the frequency-selectivity of the equivalent channel because the original signal bandwidth after localized mapping is inherently narrowband (the signal bandwidth is kept the same as the original analog signal). Therefore, MRC-FDE improves significantly the NMSE performance as shown in Fig. 3 and Fig. 4.

The MMSE weight minimizes the MSE between frequency-domain signal after the FDE and that before the subcarrier mapping. If $|\hat{H}(k)|^2 \gg \Gamma^{-1}$ (i.e., the average SNR is high), the MMSE weight approaches the ZF weight. On the other hand, if $|\hat{H}(k)|^2 \ll \Gamma^{-1}$ (i.e., the average SNR is low), the MMSE weight approaches the MRC weight. As a consequence, MMSE-FDE can restore a near frequency-nonselective channel while alleviating the noise enhancement problem. Besides, using distributed mapping, frequency diversity gain is achievable and thus, MMSE-FDE with distributed mapping provides the best performance among combinations of three FDE weights and two subcarrier mappings.

The simulated and theoretical NMSE performances of analog SC-FDE are compared in Fig. 5. A fairly good agreement between the simulation and theoretical results is seen.

Figure 6 shows a one-shot observation of voice transmission using the proposed analog SC-FDE with distributed mapping and conventional DSB when $\Gamma = 20$ dB. For conventional DSB, the frequency-nonselective fading channel with maximum Doppler frequency $f_D = 2$ Hz and ideal fast AGC are assumed. It can be clearly seen that when the channel gain h(t) drops, the noise enhancement appears in the case of conventional DSB while the received voice waveform is almost the same as the original waveform in the case of proposed analog SC-FDE. This is because the combination of MMSE-FDE and distributed mapping not only avoids the noise enhancement but also obtains significant frequency diversity gain.

Additionally, the performance comparison between analog SC-FDE and conventional DSB using space diversity is shown in Fig. 7. The number, N_t and N_r , of transmit and receive antennas are set as $N_t = 1$ and $N_r = 1 \sim 3$, respectively. At the receiver, MRC space diversity is considered.



Fig. 5 NMSE Performance comparison between simulation and theoretical results of analog SC-FDE transmission.



Fig.6 One-shot observation of voice transmission using conventional DSB and proposed analog SC-FDE.

It is shown that owing to frequency diversity gain, the performance of analog SC-FDE with distributed mapping and MMSE weight is always better than conventional DSB with MRC space diversity in case of having the same number N_r of receive antennas. When N_r increases, the space diversity gain becomes the dominant factor of performance improvement. Therefore, the performance of conventional DSB approaches that of analog SC-FDE for $N_r = 2$ and 3.

5. Conclusion

In this paper, we proposed a new analog signal transmission technique called analog SC-FDE in order to improve the 1964



Fig.7 NMSE performance comparison between analog SC-FDE and conventional DSB using space diversity in case of cosine wave transmission.

performance of analog signal transmission. In the proposed analog SC-FDE, the frequency components of analog signal are mapped over a broad bandwidth and FDE is applied to take advantage of the channel frequency-selectivity to obtain the frequency diversity gain. To evaluate the improvement in transmission performance, NMSE was introduced. We showed that the proposed analog SC-FDE achieves better NMSE performance than conventional DSB. In this paper, transmission of single analog signal stream was presented. However, it should be noted that multiple analog signal streams can be transmitted based on the principle of SC-FDMA.

In this paper, we assumed the perfect knowledge of channel state information. Analog SC-FDE performance may degrade if any practical channel estimation scheme is used. The comprehensive performance comparison between analog SC-FDE and digital transmission both with practical channel estimation scheme in different channel conditions is left as an interesting future study.

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Appendix: Derivation of σ_{ISI}^2 and σ_{noise}^2 for Analog SC-FDE

From Eq. (16), σ_{ISI}^2 can be expressed as

$$\begin{split} \sigma_{\rm ISI}^2 &= E\left[|{\rm Re} \left\{ \mu_{\rm ISI}(n) \right\}|^2 \right] = \frac{1}{2} E\left[|\mu_{\rm ISI}(n)|^2 \right] \\ &= \frac{1}{2} \times \frac{\frac{1}{M^2} \sum_{k=0}^{M-1} \sum_{k'=0}^{M-1} \tilde{H}(k) \tilde{H}^*(k')}{\left| \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \right|^2} \\ &\times \left\{ \sum_{m=0}^{M-1} \sum_{m'=0 \atop m' \neq n}^{M-1} E\left[s(m) s^*(m') \right] \exp\left[j 2\pi \left(k \frac{n-m}{M} - k' \frac{n-m'}{M} \right) \right] \right\}, \end{split}$$
(A:1)

where $n = 0 \sim M - 1$. Assuming that the band-limited signal to be transmitted has a uniform power spectrum density, we have $E[s(m)s^*(m')] = \delta(m-m')$ and then, Eq. (A·1) is rewritten as

$$\sigma_{\rm ISI}^{2} = \frac{1}{2} \times \frac{\frac{1}{M^{2}} \sum_{k=0}^{M-1} \sum_{k'=0}^{M-1} \tilde{H}(k) \tilde{H}^{*}(k')}{\left| \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \right|^{2}} \\ \times \left\{ \sum_{m=0}^{M-1} \exp\left[j2\pi \left(k - k' \right) \frac{n-m}{M} \right] - 1 \right\}.$$
 (A·2)

Using the relationship

$$\sum_{m=0}^{M-1} \exp\left[j2\pi \left(k-k'\right)\frac{n-m}{M}\right] = M\delta\left(k-k'\right), \qquad (A\cdot 3)$$

we obtain

$$\sigma_{\rm ISI}^{2} = \frac{1}{2} \cdot \frac{\frac{1}{M^{2}} \sum_{k=0}^{M-1} \sum_{k'=0}^{M-1} \tilde{H}(k) \tilde{H}^{*}(k') \{M\delta(k-k')-1\}}{\left|\frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k)\right|^{2}}$$
$$= \frac{1}{2} \cdot \frac{\frac{1}{M} \sum_{k=0}^{M-1} |\tilde{H}(k)|^{2} - \left|\frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k)\right|^{2}}{\left|\frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k)\right|^{2}}.$$
 (A·4)

Next, σ_{noise}^2 is derived. From Eq. (16), we have

$$\sigma_{\text{noise}}^{2} = E\left[|\text{Re}\left\{\mu_{\text{noise}}(n)\right\}|^{2}\right] = \frac{1}{2}E\left[|\mu_{\text{noise}}(n)|^{2}\right]$$
$$= \frac{1}{2} \cdot \frac{\frac{1}{M}\sum_{k=0}^{M-1}\sum_{k'=0}^{M-1}W(k)W^{*}(k')}{2P\left|\frac{1}{M}\sum_{k=0}^{M-1}\tilde{H}(k)\right|^{2}}$$
$$\times \left\{E\left[\hat{\Pi}(k)\hat{\Pi}^{*}(k')\right]\exp\left(j2\pi n\frac{k-k'}{M}\right)\right\}. \quad (A \cdot 5)$$

Since { $\hat{\Pi}(k)$; $k = 0 \sim M - 1$ } are the independent and identically distributed (i.i.d) zero-mean complex-valued Gaussian variables having the variance $2N_0/T_s$, we obtain

$$\sigma_{\text{noise}}^{2} = \frac{1}{2} \cdot \frac{\Gamma^{-1} \cdot \frac{1}{M} \sum_{k=0}^{M-1} |W(k)|^{2}}{\left| \frac{1}{M} \sum_{k=0}^{M-1} \tilde{H}(k) \right|^{2}}.$$
 (A·6)





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