Performance of DS-SSMA Systems with Narrow-band Interference Rejection Filters using Band-limited Power Lines.

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Abstract — An analytical method of a DS-SSMA system with a band-limited communication line is present. The calculated results of the mean bit error probability of the systems with typical power lines used as communication line is shown. From these results, it is clarified that the performance of the system is improved some by adopting a phase equalizer but is degraded largely by the attenuation of the transmission characteristics in the lower frequency band in the main lobe of the spread spectrum, and that when high level narrow-band interference noises exist on the line, an LMS adaptive prediction error filter is applicable to the rejection of these noises.

I. INTRODUCTION

The application of power lines to spread spectrum (SS) communication systems have been studied by many investigators (e.g., [1] ~ [4]). It is reported that the transmission characteristics of the system using power line as communication line vary with the change of the power loads, and that high level additive and interference noises exist on the line. Also, the frequency band-width of the system is limited to a range. The high level noise is overcome by increasing the processing gain of the system. But because of the band-limit, the information data transmission rate decreases with the increase of the gain. Furthermore, the operation of the system is disturbed by the high level interference noises, so it may become difficult to establish the communication link.

The application of the coded PN sequence having an error correcting ability to SS code sequence and the employment of the differential decoding method at the receiver to cancel the distortion on SS signal wave arising from the band-limited power line are examined, and their improvement effects are clarified [3].

In this power, we consider the performances of DS-SSMA systems using the band-limited power lines in the case where an LMS adaptive prediction error filter is used to reject L narrow-band interference noises existing on the line. In the calculations of the mean bit error probability, we use the approximation method using the characteristic function of the correlator output [6]. Where we assume that the transmission characteristic of the power line is expressed by that of a typical lamped constant circuit or a 1-pole Butterworth filter. In some calculations, furthermore we assume that by a phase equalizer at the output side of the prediction error filter, the phase characteristic of the power line is equalized ideally. In sections II and III, we present the analytical method and the calculated results.

II. SYSTEM ANALYSIS

A. The System Model

A simple model of K-users DS-SSMA system using a band-limited power (PL) line is shown in Fig.1. The k-th user uses a low pass filter (LPFk) Hk(f) in the transmitter in order to limit the transmission band to the width 2Bk. Where we assume that LPFk rejects the dc and the frequency component in the vicinity of the dc, and that the influences of these components on the performance of the system can be neglect. Let Hk(f) denote the transmission characteristic of the power line. We assume that an additive Gaussian noise nw(t) and a narrow-band in-
terference noise $n_j(t)$ exist on the line. Where $n_W(t)$ has
two-sided power spectral density $N_o/2$ and $n_j(t)$ is the
random process which we assume to be L narrow-band
Gaussian noises with total power $P_j$. The filter (IRF1)
shown in Fig.1 is used to reject the interference noise $n_j(t)$.
Now, let $(a_m^{(k)})$ and $(b_m^{(k)})$ be the SS code sequence with
period $N_c$ and the data sequence of the k-th user respectively
and $a_m^{(k)}$, $b_m^{(k)} \in \{-1,1\}$. And let $S_k(t - \tau_k)$ denote
the transmission signal of the k-th user in asynchronous
system, then

$$S_k(t - \tau_k) = \sum_{m=-\infty}^{\infty} A \cdot C_m^{(k)} \cdot g_c(t - mT_c - \tau_k) \quad (1)$$

where $A$, $\tau_k$ and $T_c$ are amplitude, random time-delay and
duration respectively, and $g_c(t) = 1$ for $0 \leq t < T_c$
and $g_c(t) = 0$ otherwise.

Furthermore, $C_m^{(k)}$ is expressed by using Gaussian symbol
$[ ]$ as follows.

$$C_m^{(k)} = a_m^{(k)} \cdot b_{m/N_c}^{(k)} \quad (2)$$

Also, we denote the transmission characteristics of the
line including the low pass filter and power line as

$$H_m^{(k)}(f) = H_m^{(k)}(f)H_L^{(k)}(f) = A_m^{(k)}(f)^2 \cdot e^{-j\theta_m^{(k)}(f)},$$

where $A_m^{(k)}(f)$ and $\theta_m^{(k)}(f)$ are the amplitude and phase
characteristics of $H_m^{(k)}(f)$ respectively.

Let $H_1^{(1)}(t)$ be the impulse response of (LPF1), then the
input signal $x(t)$ to the rejection filter (IRF1) can be written as

$$x(t) = x(t; n_1, \zeta_1; \ldots; n_K, \zeta_K)$$

$$= \sum_{k=1}^{K} \sum_{m=-\infty}^{\infty} A \cdot C_{m-n_k}^{(k)} \cdot v_k(t; m, \zeta_k)$$

$$+ \int_{-\infty}^{\infty} [n_W(u) + n_J(u)] H_1^{(1)}(t - u)du \quad (3)$$

$$v_k(t; m, \zeta_k) = T_c \int_{-\infty}^{\infty} \frac{\sin \pi fT_c}{\pi fT_c} A_m^{(k)}(f)$$

$$\cdot \cos[2\pi f(t - \frac{T_c}{2} - mT_c - \zeta_k) - \theta_m^{(k)}(f)] df$$

where we set $\tau_k = n_kT_c + \zeta_k$, $0 \leq n_k < N_c - 1$, $0 \leq \zeta_k < T_c$.

Let $H_1^{(1)}(f)$ be the frequency characteristic of IRF1,
then the characteristic of the line from the k-th user's output
to the 1-st user's correlator input, $H_k(f)$, is given by

$$H_k(f) = H_m^{(k)}(f)H_1^{(1)}(f) = A_m^{(k)}(f)e^{-j\theta_m^{(k)}(f)} \quad (4)$$

where $A_m^{(k)}(f)$ and $\theta_m^{(k)}(f)$ are the amplitude and phase
characteristics of $H_k(f)$ respectively. Then, the correlator out-
put at time $(n+1)T$, $Z_1(n)$, becomes

$$Z_1(n) = Z_1(n; n_1, \zeta_1; \ldots; n_K, \zeta_K)$$

$$= T_c \sum_{k=1}^{K} \sum_{m=-\infty}^{\infty} A \cdot C_{m-n_k}^{(k)}(n) \cdot a_1^{(1)}(n)W_k(m, \zeta_k)$$

$$+ \nu_W(n) + \nu_J(n) \quad (5)$$

$$W(m, \zeta_k) = T_c \int_{-\infty}^{\infty} \frac{\sin \pi fT_c}{\pi fT_c} A_k(f)$$

$$\cdot \cos[2\pi f(mT_c + \zeta_k) + \theta_k(f)] df$$

where $\nu_W(n)$ or $\nu_J(n)$ is expressed by

$$\nu_X(n) = \int_{-\infty}^{(n+1)T} \int_{-\infty}^{\infty} n_X(u)h_k^{(1)}(t - u)a_1(t)dudt \quad (6)$$

for $X = W$ or $J$, $h_k^{(1)}(t)$ is the impulse response of the
combined characteristic $H_k^{(1)}(f)$, $H_1^{(1)}(f)$, $a_1(t)$ is SS code
signal, and $T$ is the duration of a pulse wave of data signal
and equals $N_cT_c$.

The first term of $Z_1(n)$ in Eq.(6) includes the random
variables $b_m^{(k)}$ and $\tau_k$ for $k = 1, 2, \ldots, K$. In order to clarify
the relation between the first term and the variables,
we assume that $\tau_1 = 0$ and the intersymbol interference
arising from the line up to the input of the correlator influ-
ences the chip wave in time interval form $-MT_c$ to $MT_c$, where $M > 2$ and $2M < N_c$.

Then using $p(n_k, \zeta_k)$ as mentioned hereunder, we can show that the first term consists of the desired compo-
nent $\nu_S^{(1)}(n)$, the intersymbol interference $\nu_I^{(1)}(n)$, and the multiple-access interference $\nu_f(n)$.

Now, let $\chi_i(n_k)$ be functions of $n_k$, $i = 1, 2, \ldots, 5$, as expressed by

$$\chi_1(n_k) = 1, \quad 0 \leq n_k \leq M - 2$$

$$\chi_2(n_k) = 1, \quad 0 \leq n_k \leq M - 1$$

$$\chi_3(n_k) = 1, \quad M \leq n_k \leq N_c - M - 1$$

$$\chi_4(n_k) = 1, \quad M \leq n_k \leq N_c - M$$

$$\chi_5(n_k) = 1, \quad N_c - M + 1 \leq n_k \leq N_c - 1 \quad (7)$$

Furthermore, let $D_{k,1}(l_1, l_2; m, n_k, \zeta_k)$ be the product of a function related to the intersymbol interference,
$W(m, \zeta_k)$, and the partial cross-correlation between the
respective SS code sequences of the first and the k-th users,
as follows,

$$D_{k,1}(l_1, l_2; m, n_k, \zeta_k) = \sum_{i=1}^{l_2} a_m^{(k)} \cdot a_{n_k-m}^{(1)}W(m, \zeta_k) \quad (8)$$

where

$$W(m, \zeta_k) = T_c \int_{-\infty}^{\infty} A_k(f)$$

$$\cdot \cos[2\pi f(mT_c + \zeta_k) + \theta_k(f)] df$$

Then $p(n_k, \zeta_k)$ is expressed as

$$\rho_{-2,1}(n_k, \zeta_k) = \sum_{m=-M}^{n_k-N_c-1} D_{k,1}(N_c + m + 1, n_k)$$

$$\rho_{-1,1}(n_k, \zeta_k) = \sum_{m=-M}^{n_k-N_c-1} D_{k,1}(m + 1, n_k)$$
\[ \rho_{-1,1}(n_k, \zeta_k) = \sum_{m=-M}^{M-2} D_{k,1}(m+1, M-1) = \sum_{m=-M}^{M-2} D_{k,1}(1-M, m) + \sum_{m=-M}^{M-2} D_{k,1}(1-M, m) \]

\[ \rho_{0,1}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{0,0}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{0,2}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{0,3}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{0,4}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{0,5}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{1,1}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{1,2}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{1,3}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{1,4}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

\[ \rho_{1,5}(n_k, \zeta_k) = \sum_{m=-M}^{M-1} D_{k,1}(m+1, M-1) + \sum_{m=-M}^{M-1} D_{k,1}(1-M, m) \]

where we use \( D_{k,1}(l_1, l_2) \) instead of \( D_{k,1}(l_1, l_2; m, n_k, \zeta_k) \) to simplify the description. Using \( \chi(n_k) \) and \( \rho(n_k, \zeta_k) \), we can express \( \nu_{\alpha}^{(1)}(n), \nu_{\alpha}^{(1)}(n), \) and \( \nu_{\alpha}(n) \) as follows.

\[
\begin{align*}
\nu_{\alpha}^{(1)}(n) &= AT \cdot b_{\alpha}^{(1)}(n) \\
\nu_{\alpha}^{(1)}(n) &= AT \cdot b_{\alpha}^{(1)}(n) + AT h_{\alpha}^{(1)}(n) \\
\nu_{\alpha}(n) &= AT (n; n_2, \zeta_2; \cdots; n_K, \zeta_K) \\
&= AT \cdot \sum_{k=1}^{K} \sum_{k=1}^{K} c_{\alpha k}^{(k)} \cdot I_{\alpha}^{(k)}(n_k, \zeta_k)
\end{align*}
\]

where we put \( n_2 = \zeta_2 = 0 \), and

\[
\begin{align*}
I_{\alpha}^{(1)}(n) &= \chi_{\alpha}(n_k) \rho_{-2,1}(n_k, \zeta_k)/N_c \\
I_{\alpha}^{(1)}(n) &= \rho_{-1,1}(0, 0)/N_c \\
I_{\alpha}^{(1)}(n) &= \rho_{1,1}(0, 0)/N_c \\
I_{\alpha}^{(1)}(n) &= \chi_{\alpha}(n_k) \rho_{-2,1}(n_k, \zeta_k)/N_c \\
I_{\alpha}^{(1)}(n) &= \chi_{\alpha}(n_k) \rho_{-1,1}(n_k, \zeta_k) + \chi_{\alpha}(n_k) \rho_{-2,1}(n_k, \zeta_k) \\
&+ \chi_{\alpha}(n_k) \rho_{-1,1}(n_k, \zeta_k) + \chi_{\alpha}(n_k) \rho_{1,1}(n_k, \zeta_k)/N_c \\
I_{\alpha}^{(1)}(n) &= \sum_{i=1}^{5} \chi_{\alpha}(n_k) \rho_{0,1}(n_k, \zeta_k)/N_c
\end{align*}
\]

Eq.(5) consequently is expressed as [5]

\[ \sigma_{W}^{2} = E[\nu_{\alpha}^{(1)}(n)] = \frac{N_{c} T_{T}}{2} \int_{-\frac{1}{2} f_{s}}^{\frac{1}{2} f_{s}} \left( \frac{\sin \pi f_{T} f_{s}}{\pi f_{T} f_{s}} \right)^{2} H_{\alpha}^{(1)}(f) \cdot H_{\alpha}^{(1)}(f) \cdot S^{(1)}(f) df \]

\[ \sigma_{F}^{2} = E[\nu_{\alpha}^{(1)}(n)] = \frac{L_{\alpha} T_{T}}{2 B_{\alpha}} \int_{f_{1} \rightarrow f_{2}} \left( \frac{\sin \pi f_{T} f_{s}}{\pi f_{T} f_{s}} \right)^{2} H_{\alpha}^{(1)}(f) \cdot H_{\alpha}^{(1)}(f) \cdot S^{(1)}(f) df \]

where the total power \( P_{J} = \sum_{i=1}^{L} P_{J}^{(i)} \) and

\[ S^{(1)}(f) = 1 + 2 \sum_{q=1}^{N} \sum_{i=0}^{L} a_{i}^{(1)} a_{i+q}^{(1)} \cos(2\pi q f_{T} f_{s}) \]

By applying Eqs(7)~(15) to the approximation method as the following, we can compute the mean bit error probabilities of the systems.

B. The Prediction Error Filter

As shown in Fig.2, the filter to reject the narrow-band interference noise consists of a two-sided N-weight linear prediction error filter(IREF1) in which, according to LMS algorithm, the coefficients \( -C_{\alpha}(i) = -C_{\alpha}(i), (n = 1, 2, \cdots, N) \) are updated at the input sampling rate \( f_{s} \). Let \( x(i) \) and \( y(i) \) be the input and output signals of IREF1 at the time \( t = i/f_{s} \) respectively, then the adaptive algorithm is expressed as

\[ C_{\alpha}(i+1) = C_{\alpha}(i) + \mu y(i) [x(i - n) + x(i + n)] \]

where \( \mu = 0, \) and

\[ y(i) = x(i) - \sum_{n=1}^{N} C_{\alpha}(i) [x(i - n) + x(i + n)] \]

where \( \mu \) is the step size parameter. Put \( T_{T} = 1/f_{s} \), then the frequency characteristic of IREF1, \( H_{\alpha}^{(1)}(f) \), becomes

\[ H_{\alpha}^{(1)}(f) = 1 - 2 \sum_{n=1}^{N} C_{\alpha} \cdot \cos(2\pi n f_{T} f_{s}) \]

Fig. 2. Two-sided prediction error filter.
C. The Mean Bit Error Probability

We assume that the data symbols are mutually independent random variables and, furthermore, let \( \phi_I^{(i)}(u) \), \( \phi(u) \), and \( \phi_N(u) \) be the characteristics functions of \( v_I^{(i)}(n)/AT \), \( v_I(n)/AT \), and \( |\nu_W(n) + \nu_J(n)|/AT \) respectively, then that of \( |v_I^{(i)}(n) + \nu_I(n) + \nu_J(n)|/AT \), \( \phi(u) \), is given by

\[
\phi(u) = \phi_I^{(i)}(u) \cdot \phi_I(u) \cdot \phi_N(u) \quad (18)
\]

where

\[
\begin{align*}
\phi_I^{(i)}(u) &= \cos(u \Phi_I^{(i)}) \cdot \cos(u \Phi_I^{(i)}) \\
\phi_I(u) &= \phi_I(u; n_2, \zeta_2; \cdots; n_k, \zeta_k) \\
&= \Pi_{k=2}^{K} \left( \frac{1}{N_c A_c} \sum_{n_k=0}^{N_c-1} \int_{0}^{T} \left[ \Pi_{i=2}^{N_c} \cos[u \Phi_I^{(i)}(n_k, \zeta_k)] \right] d\zeta_k \right) \\
\phi_N(u) &= \exp\left(-\sigma^2_N/2\right) \\
\sigma^2_N &= [\sigma^2_W + \sigma^2]/(AT)^2
\end{align*}
\]

\[
(19)
\]

Using \( \phi(u) \), the bit error probability \( P_e \) for the fixed \( n_i \) and \( \zeta_i \), \( i=2, \cdots, K \) is expressed as follows.

\[
P_e = \frac{1}{2} - \frac{1}{\pi} \int_{0}^{\infty} \sin \text{uIS}_f \phi(u)du \quad (20)
\]

Therefore, the mean bit error probability \( \bar{P}_e \) is given by the average of \( P_e \) on the delay parameters \( n_i \) and \( \zeta_i \), i.e.

\[
\bar{P}_e = E_{n_i, \zeta_i}[P_e] \quad (21)
\]

III. Calculation

The transmission characteristic of a power line is varied by the loads of the line and is modeled by lumped or distributed constant circuits on the measurement results of their attenuation characteristics. These modeling have been studied by many investigators. Where, on condition that the influences of the phase characteristics of the lines can be neglected by using phase equalizers, we calculate the mean bit error probabilities of the system with the lines expressed by the typical models. As it is difficult to calculate \( \bar{P}_e \) in Eq.(21) exactly, we set the average of several \( P_e \) to \( \bar{P}_e \). Also, we assume that the interference noise \( n_j(t) \) consists of the \( L \) narrow-band interferences, i.e., \( n_j(t) = \sum_{i=1}^{L} n_j^{(i)}(t) \), and let \( f_i, B_i \) and \( P_i^{(i)} \) be the center frequency, band-width and power of the \( i \)-th interference respectively. Then the additive noise \( n_W(t) \) and interferences \( n_j^{(i)}(t), i=1, 2, \cdots, L \) can be both expressed by the sets of cosine waves with constant amplitude as follows.

\[
\begin{align*}
n_W(t) &= Aw \sum_{n=1}^{N_1} \cos(2\pi fn_W t + \alpha_n), N_1 = l_1 + M_1 \\
n_j^{(i)}(t) &= A_j^{(i)} \sum_{n=1}^{N^{(i)}} \cos(2\pi fn_j^{(i)} t + \beta_n^{(i)}), N^{(i)} = l^{(i)} + M^{(i)}
\end{align*}
\]

where \( M_1 + 1 \) and \( M^{(i)} + 1 \) are the numbers of cosine waves within the band of each noise, and both \( \alpha_n \) and \( \beta_n^{(i)} \) are random variables distributing uniformly in interval \([0, 2\pi]\).

Furthermore, \( f_s = f_B/M_1, Aw = \sqrt{2B_nN_e/(M_1 + 1)}, f_i^{(i)} = B_i/M^{(i)} \), and

\[
A_i^{(i)} = \sqrt{2P_i^{(i)}/(M^{(i)} + 1)}. \quad (21)
\]

In the design of prediction error filter, we put \( N = 20 \) and \( \mu = 10^{-5} \), and use the sampling rate \( f_s \) equaling to twice of the chip rate \( f_c \), i.e. \( f_s = 2f_c \).

![Fig. 3. The error probability \( \bar{P}_e \) against the relative band-width \( Be/Tc \) where a power line has ideal characteristic.](image)

In the calculations, we put \( M = 24 \) and \( M_1 = M^{(i)} = 500 \) for \( i = 1, \cdots, L \), and assume that Gold code sequences with \( N_c = 127 \) are used in the system, and that the relative interference band-width \( B_i/f_c \) are equal to 0.01, and that the relative center frequencies of the interference noises \( f_i/f_c \) are equal to 0.2 in case of \( L = 1 \) and to 0.2, 0.4, and 0.6 in case of \( L = 3 \) respectively, and that the ratio of the interference noise power to a user's SS signal power, \( P_i^{(i)}/A^2 \), are equal to 15dB for i = 1, 2, 3, furthermore, and that the ratio of energy for a data duration of SS signal to the power spectral density of the additive noise, \( E_o/N_o \), is constant where \( E_o = A^2T \). First, we examine the mean bit error probabilities in the case where the transmission characteristic of the power line is ideal, i.e. \( H_p(f) = 1 \) for \( |f| \leq B_c \) and \( H_p(f) = 0 \) otherwise. Fig.3 shows the variations of \( \bar{P}_e \) for the relative transmission band-width \( B_c/f_c \) and the effective behaviors of the prediction error filter used to reject the interference noises. Subsequently, we examine \( \bar{P}_e \) in case of the power line with typical transmission characteristics given by reference [3]. Fig.4 (a) and (b) shows the relation between \( \bar{P}_e \) and the relative band-width \( B_c/f_c \) in the case where a 29 inch TV is loaded into the line. As shown in Fig.4 (a) and (b), the error probability is very large in comparison with the results in case of the ideal transmission line. But the prediction error filter behaves in the same manner as the case of the ideal line.

Furthermore, by comparing the results in Fig.4 (a) with the ones in (b), it is clarified that \( \bar{P}_e \) is improved by adopting a phase equalizer. Fig.5 shows \( \bar{P}_e \) in the case where a half 3dB band-width of 1-pole Butterworth filter is 0.5f_c. As the Butterworth filter has maximally flat characteris-
Fig. 4. The error probability $P_e$ against the relative band-width $B_{eTc}$, where a 29 inch TV is loaded into a power line.

Fig. 5. The error probability $P_e$ against the relative band-width $B_{eTc}$, where a power line has 1-pole Butterworth characteristic.

IV. CONCLUSIONS

We proposed an analytical method of DS-SSMA systems with narrow-band interference noises using power lines and showed the calculated results. Also, in the design of the prediction error filter to reject the interferences, we select a half chip duration as the delay of the filter. By using this filter, the several interferences existing within the main lobe band of the spread spectrum were rejected with comparative ease. The degradation of the error probability in the case where the power line loading a 29 inch TV is used as the communication line is caused by the large attenuation of transmission characteristic of the line in the lower frequency band of the main lobe. Therefore, an adaptive channel equalizer may be necessary to the system using the power line. This problem remains as subject for further research.

REFERENCES


