OFDM Interference Suppression for DS/SS Systems Using Complex FIR Filter

Yuki SHIMIZU\(^{(a)}\) and Yukitoshi SANADA\(^{(b)}\), Members

SUMMARY In this paper, the performance of narrow band interference (NBI) rejection scheme for direct sequence spread spectrum (DS/SS) is analyzed. A 2-tapped complex FIR filter is used for filtering a chip code to suppress NBI. In this system, the spectrum of transmitted signal has a null at an arbitrary frequency. By choosing filter coefficients, the authors place this null at NBI center frequency to mitigate the effect of NBI. In this paper, an OFDM signal is considered as NBI. The performance of this scheme is theoretically analyzed by introducing Queueing model, and validated via simulation. 

key words: spread spectrum, direct sequence, narrow band interference, FIR

1. Introduction

Recently, DS/SS has achieved major popularity in wireless communication systems such as IEEE802.11b Wireless LAN, UWB, or IMT-2000 Cell Phone. Due to its large bandwidth, some of them have to share the same spectrum with narrow band systems. Although DS/SS has its own resistance against NBI, an extremely powerful interference is still a problem [1], [2].

A number of studies on NBI suppression for DS/SS have been reported. One of these approaches is to change chip code encoding [3]. Chip code interleaving and 2-tapped FIR filtering are also proposed [4]. In addition, suppression scheme with an adaptive line enhancer (ALE) is available [5].

In this paper, an FIR filter is used to modify the chip code instead of filtering the signal. Hence, this scheme does not deteriorate the performance by itself, and has the ability to mitigate NBI of an arbitrary frequency. The FIR filtering of received signal has already been shown to be useful in the presence of tone interferer by simulation [4]. This paper supposes a situation that OFDM interferer is present. It is also assumed that the interferer’s bandwidth is much narrower than that of the DS/SS system, like OFDM WLAN conflicting with DS-UWB systems of full bandwidth. The performance of the system is investigated via theoretical analysis first. Simulations are then conducted to validate the analysis.

2. System Model

2.1 System Model

Figure 1 shows a system used in this paper. The sequence \(d(n)\) is spread by filtered chip code \(c(n)\) and then transmitted at frequency \(f_c\) as \(s(t)\). During the transmission, a signal from other wireless devices \(i(t)\) is added as interference. Then AWGN \(n(t)\) is added to the received signal due to LNA at the receiver. These signals \(r(t)\) are firstly brought down to the base band. The base band signal \(b(t)\) is then sampled at the chip interval \(T_c\), and is despread to recover the sequence \(d(n)\).

2.2 Effect of FIR Filtering

In this paper, a 2-tapped FIR filter is used for spectral shaping. The filter delay time \(T_d\) is set to the chip interval \(T_c\), and the filter coefficients are \(h_0 = 1, h_1 = e^{j\theta}\). Especially, the gain response of the FIR filter has a null at a specific frequency \(f_{null}\)

\[
 f_{null} = \frac{(2p + 1)\pi + \theta}{2\pi T_c} 
\]  

(1)

where \(p\) is an integer. It is possible to shift this null to any desired frequency by choosing the phase \(\theta\) of \(h_1\).

In this system, first the original chip code \(c_o(n)\) is filtered by the FIR filter. The filter output is taken as the spreading chip code \(c(n)\). According to the gain response of the filter, the spectrum of \(c(n)\) has a null. Therefore, the signal spread by \(c(n)\) has also a spectral null at \(f_{null}\).

By setting this null to the center frequency of NBI, the spectrum friction between DS/SS and NBI can be reduced. Consequently, the effect of NBI is suppressed.

2.3 OFDM Interference

In this paper, it is assumed that OFDM interference is present. All \(N_c\) subcarriers are modulated under BPSK, and each subcarrier has amplitude \(A\) and symbol duration \(T_s = M_c \cdot T_c\). The total OFDM interference \(i(t)\) is written as

\[
i(t) = \sum_{k=1}^{N_c} A \cdot e^{j(2\pi f_c + \xi)(k-\xi)} u(t,k) 
\]  

(2)

where \(\xi\) is the phase of the whole OFDM signal. \(u(t,k) = \)
\[ h_k = e^{j\pi n T_c} \]

\[ I_c(f_d(k)) = \sum_{n=0}^{M-1} A \cdot e^{j2\pi f_d(k)nT_c} e^c(n) \]

Note that \( f_d(k) \) is frequency gap defined as \( f_d(k) = f_c - f_1 \), and \( M \) is the length of the filtered code. \( c^*(n) \) is despeading chip code which is the complex conjugate of \( c(n) \).

Since the filter coefficients are \( h_0 = 1, h_1 = e^{j\theta}, c^*(n) \) can be written as

\[ c^*(n) = h_0^* \cdot c_o(n) + h_1^* \cdot c_o(n-1) \]

\[ = c_o(n) + e^{-j\theta} c_o(n-1) \]

(4)

where \( c_o(n) \) is the original chip code. Note that \( c_o(-1) = c_o(M-1) = 0 \).

Suppose the case \( f_d(k) = f_{null} \). From Eqs. (1), (3), and (4), the absolute despread value of \( k \)-th subcarrier is obtained in the following form.

\[ I_c(f_{null}) = A \cdot \left| \left( 1 + e^{j2\pi p+1}\right) \right| \cdot e^{-j\theta} \]

\[ \cdot \left| \sum_{n=0}^{M-2} c_o(n) e^{j2\pi p+1 n+\theta(n)} \right| \]

(5)

Suppose the case no phase alternation occurs while an OFDM symbol end is located within a DS/SS bit duration. In this case, the subcarrier without phase alternation can be regarded as an unmodulated tone interference. Thus the effect of the OFDM subcarrier in this case is equal to Eq. (3).

### 3. Queuing Model for Amount of Interference

#### 3.1 Approximation

For the sake of simple analysis, approximation is conducted both upon phase and amount of despread subcarriers.

As told at Sect. 1, this paper examines a situation that the bandwidth of interfering OFDM systems is far narrower than that of victim DS/SS systems. From the victim’s point of view, frequency gap among OFDM subcarriers is little enough to be ignored. For analysis, therefore, it is possible to assume that the subcarriers are all at the same frequency \( f_1 \). This means the frequency gap \( f_d(k) \) can be written as a single variable \( f_d \). Hence after despeading, absolute value and phase of interference from any subcarrier is equal to that from another.

#### 3.2 Effect of Non-phase-shift Period

Suppose \( f_d \) is right at the spectral null \( f_{null} \). As shown at 2.3, the interference defined by Eq. (3) is totally suppressed in this case. Thus, due to assumption at 3.1, the interference of each subcarrier can be neglected when there is no OFDM symbol end within a DS/SS bit duration. According to Eq. (10) in [6], the BER of the system in this case can be written as follows.

\[ P_{ol} = \frac{1}{2} \text{erfc} \left( \sqrt{\gamma} \right) \]

(8)

Note that \( \gamma \) denotes \( \frac{E_b}{N_0} \), where \( E_b \) is the bit energy of a DS/SS signal and \( N_0 \) is the power spectrum density of AWGN, respectively.

This indicates that the interference is mainly caused by symbol alternation of OFDM subcarriers.

#### 3.3 Queuing Model for Phase-shift Period

Before introducing Queuing model, let’s examine all possible amount of interference.
First off, the jammer frequency gap $f_d$ is assumed to be at the spectral null $f_{null}$. It is also assumed that a symbol end of each subcarrier is at $m$-th sample of a DS/SS bit duration.

As explained at 2.3, interference from a single subcarrier with no phase alternation is totally rejected when $f_d = f_{null}$. Since each subcarrier is modulated under BPSK, possible interference from a single subcarrier is $I_k(f_d, m)$ (phase alternation), $-I_k(f_d, m)$ (obverse phase alternation), and $0$ (no phase alternation).

This means subcarriers with different phase alternation cancel each other. Let’s consider the case of 2 subcarriers for example. Assume the despread value of the first subcarrier is

$$I_{sc1}(f_d(1), m) = \sum_{n=0}^{M-1} c'(m, n) \cdot A \cdot e^{j2\pi f_d(1)nTc - \xi}.$$ (9)

Note that $f_d(k)$ is the frequency gap for the $k$-th subcarrier. Either subcarrier is of equal phase $\xi$ due to OFD Multiplexing. For the second subcarrier has obverse phase alternation, its despread value is

$$I_{sc2}(f_d(2), m) = \sum_{n=0}^{M-1} -c'(m, n) \cdot A \cdot e^{j2\pi f_d(2)nTc - \xi}.$$ (10)

The total interference is the sum of Eqs. (9) and (10). That is,

$$Ae^{-j\xi} \sum_{n=0}^{M-1} c'(m, n) \left( e^{j2\pi f_d(1)nTc} - e^{j2\pi f_d(2)nTc} \right)$$ (11)

According to the assumption at 3.1, $f_d(1) = f_d(2) = f_d$. Thus Eq. (11) can be considered to be 0. This is how we presume that subcarriers alternating differently should cancel each other. By these ideas, possible total interference from $N_{sc}$ subcarriers can be derived from combinations of these 3 amounts of interference.

Figure 2 shows transition of total interference. Numbers within a circle mean amount of interference normalized by $I_k(f_d, m)$. Note that the probability for interference $+I_k(f_d, m)$, $-I_k(f_d, m)$, and 0 is $p_{+1}$, $p_{-1}$, $p_0$ respectively.

Suppose the case in which $N_{sc} = 2$, for example. In this case, normalized interference from either subcarrier could be $+1$, $-1$, and 0 with probability $p_{+1}$, $p_{-1}$, and $p_0$. Thus, total interference varies from $-2$ to $+2$ as depicted at the third row (labeled as “2nd subcarrier”) in Fig. 2. Probability for specific total interference can be obtained by calculating probability for every combination which leads to that amount of interference. For instance, combinations that result in total interference 0 are as listed in Table 1.

By summing these probabilities, we obtain the probability for total interference 0 at $N_{sc} = 2$ as $(p_{+1}p_{-1} + p_0^2)$.

As the number of subcarriers grows, it becomes more and more complicated to find all combinations and calculate probability for specific amount of interference. To resolve this problem, the idea of Queueing theory is employed.

Figure 3 illustrates M/M/1 Queueing model applied in this paper. Suppose there is no people in line at the initial state. The average number of those incoming and those departing for certain time period is $\lambda$ and $\mu$, respectively. According to [7] and the Appendix, probability with which the queue length is $j$ after $N_{sc}$ periods of time, is

$$P_j(\lambda N_{sc}) = e^{-\frac{1}{2}} \frac{1}{\mu} I_1 \left( 2\lambda N_{sc} \sqrt{\frac{1}{\mu}} \right)$$ (12)

where $I_1$ is the modified Bessel function. There is obvious resemblance between interference from a single subcarrier and incoming/departing queue for a single period of time. The total queue length can be regarded as the total amount of interference. The normalized interference $+1$ from a single subcarrier is to increment the total interference, and thus

![Fig. 2 Combinations of possible amount of interference.](image)

![Fig. 3 Queue length after $N_{sc}$ periods.](image)
can be regarded as a person coming into the queue. On the other hand, the negative normalized interference $-1$ can apparently be regarded as a man getting off the queue.

As listed in Table 2, an expectation of positive interference $p_{+1} \cdot 1$ can be regarded as the average incoming queue $\lambda$. The same goes on an expectation of negative interference $p_{-1} \cdot 1$ and the average departing queue $\mu$. By this modeling, we resolve Eq. (12) into the following form.

$$P_{f}(p_{+1} \cdot N_{sc}) = e^{-(1+\frac{\mu}{p_{+1}})p_{+1} \cdot N_{sc}} \cdot \frac{p_{+1} \cdot 2 + 1}{p_{+1} \cdot 2} \cdot I_{2N_{sc}} \left( \frac{p_{+1}}{p_{-1}} \right) \cdot \left( \frac{2p_{+1}N_{sc}}{p_{+1} \cdot 2} \right). \tag{13}$$

Since each subcarrier of interference is BPSK-modulated, either kind of phase alternation occurs with equal probability $\frac{1}{2}$. Assigning $\frac{1}{2}$ to $p_{+1}$ and $p_{-1}$, we get

$$P_{f}\left( \frac{N_{sc}}{2} \right) = e^{-\frac{N_{sc}}{2}} \cdot I_{1} \left( \frac{N_{sc}}{2} \right) \tag{14}$$

as the probability distribution function for total amount of interference $\sum_{j} I_{b}(f_{d}, m)$ from $N_{sc}$ subcarriers.

Now that the probability distribution for total interference is ready, a bit error rate of the system is easily acquired in the following way.

1. Calculate BER against specific amount of interference
2. Average BERs above using Eq. (12) to obtain total BER

Apparently, possible amount of interference from all $N_{sc}$ subcarriers is

$$j \cdot I_{b}(f_{d}, m) \quad (-N_{sc} \leq j \leq N_{sc}). \tag{15}$$

According to [6], the error rate of the system against amount of interference $\sum_{j} I_{b}(f_{d}, m)$ is

$$P_{e}(m, j) = \frac{1}{2} e^{\gamma} \text{erfc} \left( \sqrt{\gamma} \right) + e^{-\gamma} \sum_{q=1}^{\infty} \frac{H_{q-1}(\sqrt{\gamma}) e^{\gamma} \cdot j \cdot I_{b}(f_{d}, m)}{2^{2q} q!} \left( \frac{2q}{\sqrt{E_{b}}} \right)$$

where $H_{q}$ is the Hermite polynomial.

Using Eq. (14) to average this probability on $j$, we get

$$P_{e}(N_{sc}, m) = \sum_{j=-N_{sc}}^{N_{sc}} P_{f} \left( \frac{N_{sc}}{4} \right) P_{e}(m, j). \tag{17}$$

Since phase alternation of OFDM symbols can occur at any sample of the DS/SS bit interval, we also need to average Eq. (17) on $m$.

$$P_{e}(N_{sc}) = \frac{1}{M-1} \sum_{m=1}^{M} \sum_{j=-N_{sc}}^{N_{sc}} P_{f} \left( \frac{N_{sc}}{4} \right) P_{e}(m, j). \tag{18}$$

### Table 2 Corresponding parameters.

<table>
<thead>
<tr>
<th>Combination</th>
<th>Queuing model</th>
</tr>
</thead>
<tbody>
<tr>
<td>$p_{+1} \cdot 1$</td>
<td>$\lambda$</td>
</tr>
<tr>
<td>$p_{-1} \cdot 1$</td>
<td>$\mu$</td>
</tr>
</tbody>
</table>

Probability with which a symbol end of interference is present within a DS/SS bit duration is $\frac{M-1}{M}$. With this probability and Eqs. (8) and (18), we finally get the BER of the system as

$$P_{\text{OFDM}}(N_{sc}) = \left( 1 - \frac{M-1}{M} \right) P_{e} + \frac{M-1}{M} P_{e}(N_{sc}). \tag{19}$$

Notice that Eq. (19) is valid only when the jammer frequency $f_{t}$ is equal to the spectral null $f_{\text{null}}$. $f_{t} \neq f_{\text{null}}$ means that interference from a subcarrier without phase shift is not suppressed completely. In such situation, the combination model and the corresponding Queuing model in Figs. 2 and 3 are no longer suitable.

### 4. Numerical Results

#### 4.1 Probability by Queuing Model

Figure 4 shows probability distribution for total amount of interference from $N_{sc} = 52$ subcarriers. The X axis means the total amount of interference normalized by $I_{b}(f_{d}, m)$. For example, 52 at X axis means that all 52 subcarriers have the same phase alternation, and cause amount of interference $52 \cdot I_{b}(f_{d}, m)$ as a whole. The Y axis means the probability for each amount of interference.

The legend “Combination” means the probability obtained by calculating amount of interference for all combinations available. The legend “Queuing model” represents the probability introduced as Eq. (14). These 2 legends agree well.

#### 4.2 Effect of Code Filtering

Table 3 shows the simulation parameters used to validate the results of analysis. The 1st and the 2nd modulation of the signal is BPSK and DS/SS, respectively. The bit rate of the DS/SS system is set to 150 Mbps, and the chip rate...
Table 3: Simulation parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>1st Mod.</td>
<td>BPSK</td>
</tr>
<tr>
<td>2nd Mod.</td>
<td>DS/SS</td>
</tr>
<tr>
<td>Bit rate of DS/SS $R_b$</td>
<td>150 Mbps</td>
</tr>
<tr>
<td>Chip rate of DS/SS $R_c$</td>
<td>3.75 Gcps</td>
</tr>
<tr>
<td>Chip code $c(n)$</td>
<td>Filtered DS-UWB</td>
</tr>
<tr>
<td></td>
<td>Ternary code (M:25)</td>
</tr>
<tr>
<td>Filter delay $T_d$</td>
<td>0.27 ns</td>
</tr>
<tr>
<td>Spectral null of DS/SS signal</td>
<td>0 GHz ($p = 0$)</td>
</tr>
<tr>
<td>Num. of OFDM subcarriers $N_w$</td>
<td>52</td>
</tr>
<tr>
<td>Total OFDM bandwidth</td>
<td>414 MHz</td>
</tr>
<tr>
<td>OFDM symbol duration</td>
<td>13 µs</td>
</tr>
<tr>
<td>OFDM subcarrier bandwidth</td>
<td>15.6 MHz</td>
</tr>
<tr>
<td>Signal-to-Interference ratio $R$</td>
<td>18</td>
</tr>
</tbody>
</table>

Fig. 5: Performance against OFDM interference.

is 3.75 Gcps. The OFDM interferer consists of 52 subcarriers. Each subcarrier is modulated under BPSK, and occupies bandwidth of 15.6 MHz. The signal’s bandwidth ratio to that of the interferer is defined as $R$. The SIR in this paper is defined as the following.

$$SIR = \frac{E_b}{N_0} \cdot A^2 \cdot T_b.$$  \hspace{1cm} (20)

Note that $T_b$ is the bit duration of DS/SS signal.

Figure 5 shows the performance of the DS/SS system in the presence of the OFDM interferer. The center frequency and the bandwidth of the interferer is $f_{null}$ and 414 MHz, respectively. The legends “Sim(Plain)...” show the simulated performance of DS/SS without any interference rejection. The rest of the legends means the performance of the proposed system obtained through simulation (labeled as “Sim...”) or via analysis (“Theory...”). The result shows code filtering is effective against OFDM interference, and the analysis before is correct in this situation.

From Figs. 4 and 5, the introduction of Queuing model can be considered appropriate when $f_I = f_{null}$.

Now the bound of the queuing model is examined. As told at Sect. 3.1, the introduction of Queuing model is mainly based on frequency approximation. Thus it is reasonable to try its validity by widening frequency interval between each subcarrier. Note that in Figs. 6 and 7, legends “Sim...” and “Theory...” means numerical and theoretical results, respectively.

Figure 6 shows how Queuing model becomes inappropriate for error rate analysis when the SIR is set to $-8$ dB. A variety of $R$ are evaluated here by changing symbol duration of interference. The parameters except OFDM interference bandwidth, OFDM symbol duration, subcarrier bandwidth, and $R$ are equal to those listed in Table 3.

As frequency interval of the interference goes greater, the gap between numerical and theoretical results becomes significant. This is clearly because our approximation at Sects. 3.1 and 3.2 is no longer appropriate, due to the growing frequency interval.

Figure 7 depicts how the theoretical BER loses its validity when the interference frequency is not at the spectral null. The X axis means frequency gap between the spectral null and the center frequency of the interference. For the sake of simple scaling, this gap is normalized by the bandwidth of the DS/SS system. $E_b/N_0$ and SIR is set to 7 dB and 0 dB, respectively. The rest of the parameters used is listed in Table 3.
As explained at Sect. 3.3, the gap between \( f_l \) and \( f_{null} \) causes incomplete suppression against interference without phase shift. This leads to far more complex combinations of interference, and eventually makes the queuing model impractical. This phenomena is obvious especially when the normalized freqency gap is about \( \pm 0.25 \).

Apparently, the numerical and the theoretical error rates stay close at the normalized gap of \( \pm 0.5 \). It is because the power spectrum density of the original chip code is 0 at those frequencies. This spectrum shape allows the system to diminish interference with no phase shift at the normalized gap of \( \pm 0.5 \). Therefore, total interference which the system may encounter is still as shown in Fig. 2 with or without code filtering. This is why these two legends are almost equal in such situations.

5. Conclusion

In this paper, the effect of code filtering against OFDM interference has been evaluated. The queuing model has been introduced for OFDM interference analysis. The theoretical BER of the system has been obtained from the queuing model, and validated with the simulations. The results have proved that the introduction of the queuing model is appropriate, and the code filtering is effective against OFDM interference.

References


Appendix: Derivation of Eq. (12)

Let \( \lambda \) and \( \mu \) denote average incoming queue and departing queue, respectively. The probability with which the resultant queue length is \( j \) after \( l + \Delta l \) time periods is given by

\[
p_j(l + \Delta l) = p_j(l) - \lambda \Delta l - \mu \Delta l + p_{j-1}(l + \mu \Delta l)
\]

\[
= p_{j-1}(l) \lambda \Delta l + \sum_i o_i(\Delta l)
\]  

(A-1)

where \( \sum_i o_i(\Delta l) \) is the term which includes the all terms with \( \lambda \mu \) and it turns to be 0 when \( \Delta l \rightarrow 0 \). Note that unlike usual Queueing model, queue length \( j \) could be either positive or negative in this paper.

Considering that \( \Delta l \rightarrow 0 \), we obtain

\[
\frac{d}{dl} p_j(l) = \lambda p_{j-1}(l) + \mu p_{j+1}(l) - (\lambda + \mu) p_j(l).
\]  

(A-2)

Suppose the generating function of \( p_j(l) \) is

\[
G'(z, \tau) = \sum_{j=\infty}^\infty p_j(\tau) z^j
\]  

(A-3)

where

\[
\tau = \lambda l.
\]  

(A-4)

From Eqs. (A-2) and (A-3), the generating function can be written as

\[
\frac{\partial G^*}{\partial \tau} = \left( \rho z - (1 + \rho) + \frac{1}{z} \right) G^*
\]  

(A-5)

where

\[
\rho = \frac{\lambda}{\mu}.
\]  

(A-6)

Assuming that \( G^* \) contains the terms with \( z \) amd \( \tau \), and those terms can be separated, we get

\[
G^*(z, \tau) = \psi(z) \psi(\tau).
\]  

(A-7)

From Eqs. (A-5) and (A-7), \( \psi \) can be expressed as

\[
\psi(\tau) = e^{-(1+\rho)\tau} \sum_{j=\infty}^\infty I_j(2\tau \sqrt{\rho} z) z^j.
\]  

(A-8)

On the other hand, let \( \varphi \) denote by series as

\[
\varphi(z) = \sum_{j=\infty}^\infty \alpha_j z^j
\]

\[
= \sum_{j=\infty}^\infty \alpha_j z^j.
\]  

(A-9)

From Eqs. (A-7), (A-8), and (A-9), the generating function is

\[
G^*(z, \tau) = e^{-(1+\rho)\tau} \sum_{j=\infty}^\infty \frac{1}{z^j} \alpha_j
\]

\[
\times \sum_{j=\infty}^\infty I_j(2\tau \sqrt{\rho}) z^j.\]

(A-10)

From Eqs. (A-3) and (A-10), \( p_j \) can be written as

\[
p_j(\tau) = e^{-(1+\rho)\tau}
\]

\[
\times \sum_{j=\infty}^\infty \alpha_j \rho^{\mu j} I_{j+1}(2\tau \sqrt{\rho}).
\]  

(A-11)

As the initial queue length at \( l = 0 \) is 0, the boundary conditions are
\[ p_j(0) = \begin{cases} 1 & (j = 0) \\ 0 & (j \neq 0) \end{cases} \]  \hspace{1cm} (A·12)

and

\[ I_j(0) = \begin{cases} 1 & (j = 0) \\ 0 & (j \neq 0) \end{cases} \]  \hspace{1cm} (A·13)

From Eqs. (A·11) and (A·12), we obtain

\[ \alpha_j = \begin{cases} 1 & (j = 0) \\ 0 & (j \neq 0) \end{cases} \]  \hspace{1cm} (A·14)

From Eqs. (A·11), (A·12), (A·13), and (A·14), the probability with which the resultant queue length is \( j \) after \( l \) time periods is given by

\[ P_j(\tau) = e^{-\tau(1+\rho_\tau)p^*}I_j(2\sqrt{\tau}). \]  \hspace{1cm} (A·15)

Yuki Shimizu was born in Kyoto in 1982. He received his B.E. degree in electronics engineering from Keio University, Yokohama Japan in 2006. Since April 2006, he has been a graduate student in School of Integrated Design Engineering, Graduate School of Science and Technology, Keio University. His current research interest is in DS/SS communication systems.

Yukitoshi Sanada was born in Tokyo in 1969. He received his B.E. degree in electrical engineering from Keio University, Yokohama Japan, his M.A.Sc. degree in electrical engineering from the University of Victoria, B.C., Canada, and his Ph.D. degree in electrical engineering from Keio University, Yokohama Japan, in 1992, 1995, and 1997, respectively. In 1997 he joined the Faculty of Engineering, Tokyo Institute of Technology as a Research Associate. In 2000 he joined Advanced Telecommunication Laboratory, Sony Computer Science Laboratories, Inc, as an associate researcher. In 2001 he jointed Faculty of Science and Engineering, Keio University, where he is now an assistant professor. He received the Young Engineer Award from IEICE Japan in 1997. His current research interest is in software defined radio and ultra wideband systems.