Synchronization algorithm of asynchronously captured data packets and instantaneous power-line frequency data for PLC analysis

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Abstract: The instantaneous power-line frequency synchronized superimposed chart was proposed by the authors to visualize the effects of an AC adapter for PLC (Power Line Communications) transmission. In order to most properly draw the superimposed chart, it is requisite to synchronize data packets and instantaneous power-line frequency data, which are asynchronously captured by separate PCs (Personal Computers). This paper proposes a synchronization algorithm for these data, and shows the effectiveness of the proposed algorithm. Furthermore, an algorithm to compensate the fixed frequency measurement error made by a digital multimeter is proposed. Finally, it is shown that the communications forbidden time, the duration when PLC burst signals are not correctly received due to the effects of the AC adapter, is properly shown by adopting the proposed two algorithms as well as how burst signals containing different number of data packets are transmitted in the moments other than the communications forbidden time.

Keywords: power line communications, data synchronization algorithm

Classification: Transmission Systems and Transmission Equipment for Communications
References


1 Introduction

A PLC system employs power-lines as transmission lines, and transmits burst signals containing multiple data packets with high throughput. Some AC adapters of mobile phones affect transmission of the PLC signals. It was reported that two kinds of transfer functions periodically appear, synchronized with a half-cycle of the power-line frequency [1, 2]. It was found by the authors that the communications forbidden time, the duration when PLC burst signals are not correctly received, appears once in a half-cycle of the power-line frequency due to the effects of the AC adapter [3]. A chart named the instantaneous power-line frequency synchronized superimposed chart, referred to as the superimposed chart, was also proposed by the authors [4] to visualize the communications forbidden time as well as how burst signals containing different number of data packets are transmitted in the moments other than the forbidden time. Note that the objective of the superimposed chart is to illustrate the actual computer communications through the PLC transmission affected by the AC adapter. To draw the superimposed chart properly, it is requisite to synchronize asynchronously captured data packets and instantaneous power-line frequency data. This paper proposes a synchronization algorithm for these data, and shows the effectiveness of the proposed algorithm.
2 Synchronization algorithm for asynchronously captured data packets and instantaneous power-line frequency data

2.1 Measurement network configuration

Fig. 1(a) depicts the measurement network configuration. UDP (User Datagram Protocol) data packets with a rate of 30 Mbit/s are generated by data packet generating software [5], supplied to the Tx (Transmit) HD (High-Definition)-PLC adapter [6], transmitted in a form of a burst signal, nominally containing 5 or 6 data packets, and received by the Rx (Receive) PLC adapter. The data packets
received by a NIC (Network Interface Card) of the Rx PC are captured by data packet capturing software [7] with timestamps indicating the completion time of the capturing process for each data packet. Instantaneous power-line frequency data are measured using a digital multi-meter (DMM), and stored in a separate PC with timestamps. The DMM [8] in Fig. 1(a) periodically outputs the instantaneous power-line frequency data every 200 msec. Two data measurements are manually initiated with timestamps accumulatively increased from the time 0, and are asynchronous mainly due to the processing delay in the DMM. Hence, the timestamps of these data should be adjusted to be synchronous each other to draw the superimposed chart properly.

2.2 Superimposed chart
The instantaneous power-line frequency synchronized superimposed chart [4] is defined by a two-dimensional chart illustrating all the burst signals, whose timestamps are represented by the timestamp of the forefront data packet of each burst signal, where the x coordinate of the burst signal is given by the remainder of the timestamp divided by the half-period of the instantaneous power-line frequency, and the y coordinate by the timestamp for that burst signal. Figs. 1(b) and (d) show the superimposed charts before synchronizing the burst signals with the power-line frequency data shown in Fig. 1(c). The white belt in the center, where no burst signals are received, is named as the communications forbidden time, which is winding alone the y axis due to the asynchronous nature of the captured data. The burst signals are classified in accordance with the number of data packets contained in a burst signal. It is observed from Fig. 1(d) that the number of data packets contained in a burst signal is reduced in the time region reaching the forbidden time, and always becomes one at the moments just adjacent to the commencing moment of the forbidden time. The burst signals containing a single data packet will be referred to as the BSSPs hereafter.

2.3 Synchronization algorithm
This paper proposes a synchronization algorithm, assuming that the BSSPs just adjacent to the commencing moment of the forbidden time align perpendicular to the y axis, when these data are synchronized. This assumption is from the fact that the commencing phase of the charge current of the AC adapter is fixed within the half-cycle of the power-line frequency, whereas the terminating phase varies, which was verified by the circuit simulation, and it is considered that the forbidden time is closely related with the charge current.

In the proposed algorithm, these data are synchronized by adding the correction time $\Delta t$ to the timestamps of the instantaneous power-line frequency data. Specifically, the correction time $\Delta t$ is incremented from a negative lower limit to an upper limit by a unit time, defined by the data output interval of the DMM, and an optimum correction time is searched. The detailed flows are as follows.

(STEP 1) Set an initial value of the correction time $\Delta t$ to the negative lower limit, and add the correction time $\Delta t$ to the timestamps of the instantaneous power-line frequency data.
(STEP 2) Extract only the BSSPs from all the burst signals.

(STEP 3) Draw the superimposed chart with the BSSPs and the corrected instantaneous power-line frequency data, as shown in Fig. 2(a).

(STEP 4) Keep only the BSSPs on a thick line, and remove the other BSSPs manually, as shown in Fig. 2(b).

(STEP 5) Assign the identification $i$ sequentially with the initial value of 1 on the remaining BSSPs.

(STEP 6) Calculate a regression line of the BSSPs with the least-square method [9] in the form of Eq. (1), where $a$ is a slope of the regression line and $b$ is a y-intercept. Next, calculate $\text{distance}(i)$, the distance between the $i_{th}$ BSSP and the regression line, by Eq. (2), where $x_i$ and $y_i$ are the x and y coordinates of the $i_{th}$ BSSP within the superimposed chart, as shown in Fig. 2(c).

$$y = ax + b \quad (1)$$

$$\text{distance}(i) = \frac{|ax_i - y_i + b|}{\sqrt{1 + a^2}} \quad (2)$$

(STEP 7) Add 1 to $i$. If $i$ exceeded $n$, where $n$ is the total number of the BSSPs, go to STEP 8. Otherwise, go back to STEP 6.

(STEP 8) Calculate $\text{avg.distance}(j)$, the average distance for all the distance data, by Eq. (3), where $j$ is the identification with the initial value of 1, corresponding to a specific correction time.

$$\text{avg.distance}(j) = \frac{\sum_{i=1}^{n} \text{distance}(i)}{n} \quad (3)$$

![Supplementary charts for the proposed synchronization algorithm.](image-url)
\[
\text{avg.distance}(j) = \frac{1}{n} \sum_{i=1}^{n} \text{distance}(i)
\]  

(3)

(STEP 9) Add 1 to \( j \). Add the unit time, the interval time of the instantaneous power-line frequency data output by the DMM, to the correction time \( \Delta t \), add the renewed correction time \( \Delta t \) to the timestamps of the instantaneous power-line frequency data, and return to STEP 3. If the correction time \( \Delta t \) exceeded the upper limit, go to STEP 10.

(STEP 10) Determine the correction time \( \Delta t \) which minimizes \( \text{avg.distance}(j) \) as the optimum correction time, and finish the algorithm.

3 Results and discussions

The data packets and instantaneous power-line frequency data were applied for the synchronization algorithm. The unit time in STEP 9 was set to 200 msec, and the lower and upper limits were set to \(-3.0\) and \(0.0\) sec, respectively. Fig. 3(a) shows the calculated average distance as a function of the correction time \( \Delta t \). The burst signals just adjacent to the commencing moment of the forbidden time align most linearly when the average distance takes the minimum and optimum value of \(-2.2\) sec. This time difference is mainly due to the processing delay of the DMM, including the error made by the manually and asynchronously initiated measurements.

The DMM generates a measurement error as specified in the specifications [8]. The error was assumed to be fixed during the measurements, and searched among the error range specified in the specifications by compensating specific fixed errors and by observing the resultant superimposed charts. In the example calculation using the data with the optimum \( \Delta t \) of \(-2.2\) sec, the error was determined to be \(-8.3 \times 10^{-4}\) Hz. The superimposed chart after compensating the measurement error is shown in Fig. 3(b). The forbidden time is shown as a white belt perpendicular to the horizontal axis, and how burst signals containing different number of data packets are transmitted in the moments other than the forbidden time is clearly illustrated. The white forbidden time belt is strongly correlated with the purple belt, where the number of data packets is typically 11 or 12, and the commencing moments of these two belts are separated by 2.26 msec, which is the typical burst signal interval. By comparing Fig. 1(b) and Fig. 3(b), the effectiveness of the proposed two algorithms is clarified as explained above.

It is also verified that the BSSPs always appear at the moments just before the commencing moment of the forbidden time, as long as the electric appliances generate the forbidden time, those include some of LED lamps, electric fans, and portable game players and so on.
4 Conclusions

The synchronization algorithm for the asynchronously captured data packets and instantaneous power-line frequency data is proposed. The example calculation showed the effectiveness of the proposed algorithm. Furthermore, the fixed frequency measurement error of the DMM was searched, determined and compensated, and the resultant superimposed chart properly showed the communications forbidden time belt as well as how burst signals containing different number of data packets are transmitted in the moments other than the forbidden time.

Fig. 3. Example calculation results using the proposed algorithms.
Experimental validation of communication disturbance observer for networked control systems with information losses

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Abstract: Networked control systems (NCSs) using the Internet will soon become widespread. In NCSs, time delays and information losses cause system performance degradation and destabilization. It is known that a communication disturbance observer (CDOB)-based compensator can estimate and suppress the effect of time delays on the system as a network disturbance. We applied a CDOB-based compensator to NCSs with not only time delays but also information losses and theoretically analyzed the information-loss compensation scheme. However, the performance of the CDOB-based compensator for only information losses in real NCSs has not been experimentally evaluated. This letter demonstrates an information-loss compensation scheme using a CDOB for NCSs. Experiments using a networked motion control system show that the CDOB-based compensator can estimate and suppress the effect of information losses on an NCS.

Keywords: sensor-actuator network, machine-to-machine, motion control, networked control system, information loss

Classification: Network

References


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1 Introduction

In recent years, the rapid spread of high-performance computing technologies and broadband communication networks has accelerated the realization of various networked control systems (NCSs) [1, 2]. NCSs are categorized into “control of network” and “control over network.” As one of the “control over network” technologies, networked motion control is currently an attractive research topic. For example, teleoperation robots, which are utilized in space, under water, or in nuclear power plants, have been developed based on networked motion control technologies.

In NCSs, the existence of time delays and information losses in the feedback control loop is a significant problem in the design of an appropriate controller. The Smith predictor (SP) is a classical time-delay compensator using a time-delay model [3]. The SP needs a precise time-delay model, and the difference between the model and actual time delay causes system performance degradation and destabilization. The adaptive Smith predictor (ASP) measures time delays and updates the time-delay model to remove the modeling error of time delays [4]. Natori et al. [5, 6] proposed a communication disturbance observer (CDOB)-based compensator to estimate and suppress the effect of time delays on NCSs without any time-delay models. In addition, the combination of a CDOB and jitter buffers improved the performance of time-delay compensation scheme [7].

Since the number of NCSs using imperfect networks, such as the Internet and wireless networks, are expected to increase in the future, an effective compensator of information losses to build more dependable networked motion control systems is required. However, the behavior of conventional time-delay compensators, i.e., the SP, ASP, and CDOB, when information losses occur has not been analyzed. The authors previously applied the conventional time-delay compensators to NCSs with not only time delays but also information losses [8]. In [8], it was theoretically shown that only a CDOB can compensate the effect of the information losses on
NCSs. Codrean et al. [9] utilized the CDOB in the experiments, where there are time delays and information losses. However, the effects of time delays and information losses in NCSs with the CDOB have not been evaluated separately. It is very important to separately discuss the effects of information losses and time delays when we design an appropriate network for NCSs.

This letter briefly describes the concept of the ND including information losses proposed in [8] and experimentally evaluates the CDOB-based compensator in an NCS when only information losses exist. Experiments using a networked position control system, which comprises a rotary motor and virtual lossy networks, show that the CDOB-based compensator can estimate and suppress the effect of the information losses on the NCS.

2 Networked motion control

This section describes a networked position control system comprising a motor and a robust control technique using a disturbance observer (DOB).

2.1 Position control system

The networked position control system of a motor comprises the controller $G_C$, remote system $G_P$, and network elements, as shown in Fig. 1. In this figure, $x^{cmd}$, $x^{res}$, and $f^{ref}$ are the position command, position response, and force reference, respectively. The subscript $d$ denotes the value after passing through the network element. The remote system includes a DOB to realize robust motion control. The network elements include time delays, $T_1$ and $T_2$, and information losses $L_1$ and $L_2$. The value of each of the parameters $L_1$ and $L_2$ is stochastically 0 or 1.

As one of the position controllers, a proportional-derivative (PD) controller can be implemented as

$$G_C = J_n(K_p + K_ds),$$

where $J_n$, $K_p$, $K_d$, and $s$ denote a moment-of-inertia model of the motor, a proportional gain, a derivative gain, and the Laplace operator, respectively.

The transfer function of the total control system shown in Fig. 1 is expressed as

$$\frac{x^{res}}{x^{cmd}} = \frac{G_C G_P L_1 e^{-T_1s}}{1 + G_C G_P L e^{-T_2s}},$$

Fig. 1. Networked position control system with DOB.
where \( T = T_1 + T_2 \) and \( L = L_1L_2 \). The denominator of the transfer function includes time-delay and information-loss elements. The network elements must be considered in the design of the controller \( G_C \).

2.2 DOB

Motion control systems include various uncertainties and disturbances, e.g., the parameter variation of moment of inertia, \( \Delta J \), the parameter variation of torque constant, \( \Delta K_t \), and external force, \( f_{ext} \). In this study, a DOB was implemented to suppress uncertainties and disturbances. The DOB can estimate and compensate the uncertainties and disturbances as a disturbance force \( f_{dis} \). The disturbance force estimated by the DOB, \( \hat{f}_{dis} \), is expressed as

\[
\hat{f}_{dis} = \frac{g_{dob}}{s + g_{dob}} f_{dis},
\]

where \( g_{dob} \) is the cut-off frequency of the low-pass filter (LPF). The DOB suppresses the disturbance \( f_{dis} \) ideally if the cut-off frequency \( g_{dob} \) is sufficiently large. In this study, we assumed the remote system \( G_P \) with the DOB is given as \( G_P = \frac{1}{L_1} \).

3 Information-loss compensation using CDOB

This section describes the concept of a network disturbance (ND) and a CDOB-based compensator for NCSs with time delays and information losses. A block diagram of the networked position control system with the CDOB is shown in Fig. 2.

![Fig. 2. Networked position control system with CDOB.](image)

The CDOB-based compensator can estimate and suppress the effect of time delays and information losses on the system as the ND \( f_{nd} \). The ND is expressed as

\[
f_{nd} = (1 - L e^{-T_1}) f_{ref}.
\]

When the remote system model \( \hat{G}_P \) is equal to the actual remote system \( G_P \), the output of the CDOB, \( x_{cmp} \), is calculated as

\[
x_{cmp} = \frac{G_{cdob}}{s + g_{cdob}} G_P f_{nd},
\]

where \( g_{cdob} \) denotes the cut-off frequency of the LPF. The effect of the ND is estimated by the compensation signal \( x_{cmp} \), if the cut-off frequency \( g_{cdob} \) is
sufficiently large. In this study, a first-order LPF was used as the first demonstration. The LPF is generally designed on the basis of the ND characteristics.

The transfer function of the total control system shown in Fig. 2 is expressed as

$$\frac{x^{\text{err}}}{x^{\text{cmd}}} = \frac{G_C G_P L_1 e^{-Ts}}{1 + G_C G_P}.$$ (6)

The denominator of the transfer function does not include any time-delay and information-loss elements. The controller $G_C$ can be designed without considering the network elements.

4 Experiments

In this section, the experimental system and experimental results are presented to verify the CDOB-based compensation scheme under information losses.

4.1 Setup

In the experiments, as shown in Fig. 3(a), the angular position of the rotary motor was controlled by the PD controller with the DOB over the virtual network with information losses. For example, in wireless NCSs, the controller is usually implemented on the base station side. The assumption of information losses on the feedback path is reasonable, since packet losses occur mainly in the upstream direction.

The moment-of-inertia model $J_n$, proportional gain $K_p$, derivative gain $K_d$, cut-off frequency of the DOB $g_{dob}$, and cut-off frequency of the CDOB $g_{cdob}$ were set to 0.0166 kg·m², 400, 40, 100 rad/s, and 100 rad/s, respectively. The feedback gains were determined such that they provided a critical damping response. The sampling period of the control system was set to 1 ms. The time delays were not considered to verify the performance of information-loss compensation, i.e., $T_1 = T_2 = 0$ ms. Uniformly distributed information losses were emulated on the feedback path with a loss rate of 90% or 99%, i.e., $L_2$ was set to 0 at a frequency of the loss rate, and $L_1$ was always set to 1.

4.2 Results

The results of experiments without the CDOB are shown in Figs. 3(b) and 3(c), and the results with the CDOB are shown in Figs. 3(d) and 3(e). Figs. 3(b) and 3(d) show the results when the loss rate was set to 90% and Figs. 3(c) and 3(e) show the results when it was set to 99%. The results without the CDOB did not converge to the command when the loss rate was 99%, whereas they did when the loss rate was 90%. The results with the CDOB converged to the command with some steady-state errors independently of the loss rate. The steady-state errors are caused by the effect of the modeling error between the remote system model $\hat{G}_P$ and the actual remote system $G_P$, and the filter design of the CDOB. It was confirmed that the CDOB could compensate the effect of information losses in real motion control systems while generating steady-state errors.
5 Conclusion

This letter demonstrated an information-loss compensation scheme using a CDOB for networked motion control systems. The experiments using the networked position control system of a rotary motor showed that the CDOB-based compensator could compensate the effect of the information losses while generating steady-state errors.

Our further study includes the application of the CDOB-based compensation scheme to NCSs over the Internet and the consideration of the time-delay and information-loss characteristics of various network systems in the design of the LPFs of the DOB and CDOB to reduce the steady-state errors.

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Bi-generalized space shift keying over MIMO channels

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Abstract: Bi-space shift keying (BiSSK) can increase the spectral efficiency of space shift keying (SSK) but requiring a large number of available transmit antennas. A new SSK-based modulation scheme is proposed which requires a smaller number of transmit antennas than what required in BiSSK to deliver the same transmission rate at a negligible performance loss. The proposed scheme is obtained by multiplexing two in-phase and quadrature generalized SSK (GSSK) streams and optimizing the carrier signals transmitted by the activated antennas. Performance of the proposed scheme is compared with other SSK-based schemes via minimum Euclidean distance analysis and computer simulation.

Keywords: MIMO channels, spatial modulation, space shift keying

Classification: Wireless Communication Technologies

References


1 Introduction

In a MIMO system, when data is simultaneously transmitted on the same frequency band from multiple transmit antennas, inter-channel interference (ICI) exits. To completely eliminate ICI, [1] proposed a technique, called spatial modulation (SM), in which only one antenna is activated at any transmission time. With this strategy, the antenna index also involves in the process of sending data to the receiver. To further reduce the decoding complexity, amplitude/phase modulation symbols are removed and data is solely conveyed by antenna indices. This method is referred to as SSK [2]. Although having very low decoding complexity, the drawback of SSK is low spectral efficiency. There are many recent studies on increasing the spectral efficiency of SSK. For example, the generalized SSK (GSSK) [3] activates more than one antenna to yield a larger number of codewords. The spectral efficiency of GSSK is $\log_2(\frac{N_t}{n_t})$ bits/sec/Hz, where $N_t$ is the number of available transmit antennas while $n_t$ is the number of activated antennas. BiSSK [4] is another scheme that doubles the rate of SSK by multiplexing two orthogonal channels: in-phase and quadrature channels. The transmission on each channel is determined by the SSK rule. Recently, another scheme that also uses the multiplexing technique to enhance the spectral efficiency is layered-SSK (LSSK) [5]. In LSSK, the layers are determined by the SSK rule and multiplexing is performed by adding/dropping active antennas.

The new scheme proposed in this paper, termed bi-generalized SSK (BiGSSK), employs GSKK rule for each in-phase/quadrature data stream. As such, BiGSSK has the same advantage of low detection complexity as with SSK, while doubling the transmission rate of GSSK. By optimizing the signals transmitted over activated antennas, it is shown that such spectral efficiency advantage of BiGSSK is compromised only by a negligible performance loss. Numerical results and performance comparison with GSSK, BiSSK, LSSK demonstrate the efficiency of the proposed BiGSSK.

2 System model and performance analysis

The baseband input/output signal model for a $N_t \times N_r$ MIMO system operating over a frequency-flat block fading channel and employing SSK-based modulation is

$$y = \sqrt{E_s}Hx + n.$$  

Here $x$ is a $N_t \times 1$ transmit symbol vector comprising of nonzero elements (corresponding to activated antennas) and zero elements (corresponding to idle antennas). The transmit symbol vector $x$ is drawn with equal probabilities from a codebook $A$ and $E(\|x\|^2) = 1$. The $N_r \times 1$ vector $y$ is the receive signal vector, while $n$ is $N_r \times 1$ noise vector whose entries are i.i.d $\mathcal{CN}(0, N_0)$ random variables. The $N_r \times N_t$ matrix $H$ is the channel matrix, whose entries are i.i.d $\mathcal{CN}(0, 1)$ random variables, i.e., the channels are flat Rayleigh fading. The constant $E_s$ is the average transmitted energy per each symbol vector.

Given the signal model in (1), the ML detection of the transmitted vector $x$ is:

$$\hat{x}_{ML} = \arg \min_{x \in A} \| y - \sqrt{E_s}Hx \|^2$$  

The bit error probability can be approximated as
where $\kappa$ is the total number of codewords, $P(x_i \rightarrow x_j)$ is the pairwise error probability (PEP) of deciding on $x_j$ given that $x_i$ was transmitted, and $N_{i,j}$ is the number of bits in error when choosing $x_j$ over $x_i$. Conditioned on $H$, the PEP is

$$P(x_i \rightarrow x_j | H) = Q\left(\frac{E_s}{2N_0} \|Hx_i - Hx_j\|^2\right) = Q\left(\sum_{i=1}^{2N_r} \epsilon_n^2\right)$$

where the $Q$-function is defined as $Q(x) = \int_x^{\infty} \frac{1}{\sqrt{2\pi}} e^{-t^2/2} dt$ and $\epsilon_n$ is a Gaussian random variable with zero mean and variance $\sigma_n^2 = \frac{E_s}{N_0} \|x_i - x_j\|^2$. It follows that the random variable $\zeta_{ij} = \sum_{n=1}^{2N_r} \epsilon_n^2$ obeys a Chi-squared distribution with $2N_r$ degrees of freedom, whose pdf is given by $f(\zeta_{ij}) = \frac{\zeta_{ij}^{N_r-1} \exp\left(-\frac{\zeta_{ij}}{2\mu_n}\right)}{(2\mu_n)^{N_r/2} \Gamma(N_r)}$, where $\Gamma(\cdot)$ is the Gamma function. The unconditioned PEP is hence obtained as

$$P(x_i \rightarrow x_j) = E[P(x_i \rightarrow x_j | H)] = \int_0^{\infty} Q(\sqrt{\zeta_{ij}}) f(\zeta_{ij}) d\zeta_{ij}$$

which has a closed-form expression as [7]:

$$P(x_i \rightarrow x_j) = \delta_{ij}^{N_r} \sum_{m=0}^{N_r} \binom{N_r - 1 + m}{m} \left(1 - \delta_{ij}\right)^m,$$

where $\delta_{ij} = \frac{1}{2} \left(1 - \sqrt{1 - \frac{4}{1 + 4\sigma_n^2}}\right)$. Substituting (6) into (3) give an upper bound of $P[\text{bit error}]$ for a SSK-based modulation scheme. It should be pointed out that, although the preceding analysis is presented for a MIMO frequency-flat fading channel, it can be readily extended to frequency-selective and/or time-selective fading scenarios. This is because the input/output models for these channels have the same form as in (1) [8] and the PEP analysis is based on ideal channel estimation.

To gain a better understanding of the error performance of a SSK-based modulation scheme, examine the following looser upper bound [9]:

$$P(x_i \rightarrow x_j) \leq \frac{1}{2} \left(\sigma_{ij}^2 + 1\right)^{-N_r} \leq a \left(\frac{E_s}{N_0}\right)^{-N_r} \left(\|x_i - x_j\|^2\right)^{-N_r}$$

where $a = 4^{N_r}/2$. The above expression clearly indicates that the error probability depends on the Euclidean distances among possible transmit symbol vectors. In the special case that each transmit symbol vector contains only 1 and 0 (e.g. in SSK or GSSK), the Euclidean distances are the same as the Hamming distances.

3 Proposed Bi-generalized space shift keying

The analysis in the previous section shows that the performance of a SSK-based modulation scheme depends on the Euclidean distance between any two codewords (i.e., two transmit symbol vectors). As discussed before, the BiSSK scheme multiplexes one real number (+1) and one imaginary number (+j) based on two input bits where the indices of the active antennas carrying those real and imaginary numbers are determined by the SSK rule. Different from BiSSK, in order to improve the spectral efficiency, the proposed BiGSSK multiplexes multiple real and imaginary numbers based on a group of $\lambda > 2$ information bits. Specifically, $\lambda/2$ bits select a subset of antennas to carry real numbers, while the other $\lambda/2$ bits
select a subset of antennas to carry imaginary numbers. The selection of antenna subsets for either real or imaginary numbers follows the GSSK rule.

To illustrate the proposed scheme, Table I-(a) describes a partial codebook for the BiGSSK scheme with \( N_t = 5 \) and \( n_t = 2 \). This means that there are \( \frac{5}{2} = 10 \) different antenna subsets, hence \( \lambda/2 = 3 \) bits can be carried by either the real or imaginary data stream in each time slot. Let \( s_1 \) and \( s_2 \) denote two bit groups, each having \( \lambda/2 \) bits. Then Table I-(a) shows codewords corresponding to \( s_1 = 000 \), while the GSSK rule to select a subset of \( n_t = 2 \) active antennas out of \( N_t = 5 \) available antennas is as follows: \( 000 \rightarrow [1, 1, 0, 0, 0]^T \), \( 001 \rightarrow [1, 0, 1, 0, 0]^T \), \( 010 \rightarrow [1, 0, 0, 0, 1]^T \), \( 011 \rightarrow [1, 0, 0, 0, 1]^T \), \( 100 \rightarrow [0, 1, 1, 0, 0]^T \), \( 101 \rightarrow [0, 1, 1, 0, 0]^T \), \( 110 \rightarrow [0, 1, 0, 1, 0]^T \), \( 111 \rightarrow [0, 0, 1, 1, 0]^T \).

**Table I.** Examples of codebooks in the proposed BiGSSK scheme.

(a) Partial codebook in BiGSSK with \( s_1 = 000 \), \( N_t = 5 \).

<table>
<thead>
<tr>
<th>Information bits ( s_1, s_2 )</th>
<th>The symbol vector ( \mathbf{x} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>000,000</td>
<td>((1 + j, 1 + j, 0, 0, 0)^T/2)</td>
</tr>
<tr>
<td>000,001</td>
<td>((1 + j, 1, j, 0, 0)^T/2)</td>
</tr>
<tr>
<td>000,010</td>
<td>((1 + j, 1, 0, j, 0)^T/2)</td>
</tr>
<tr>
<td>000,011</td>
<td>((1 + j, 1, 0, 0, j)^T/2)</td>
</tr>
<tr>
<td>000,100</td>
<td>((1, 1 + j, 0, 0, 0)^T/2)</td>
</tr>
<tr>
<td>000,101</td>
<td>((1, 1 + j, 0, 0, j)^T/2)</td>
</tr>
<tr>
<td>000,110</td>
<td>((1, 1 + j, 0, 0, j)^T/2)</td>
</tr>
<tr>
<td>000,111</td>
<td>((1, 1, j, 0, 0)^T/2)</td>
</tr>
</tbody>
</table>

(b) BiGSSK symbol mapping with \( N_t = 4 \).

<table>
<thead>
<tr>
<th>Information bits ( s_1, s_2 )</th>
<th>BiGSSK symbol vector ( \mathbf{x} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>00,00</td>
<td>([-\sqrt{3} - j, -\sqrt{3} - j, 0, 0, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>00,01</td>
<td>([-\sqrt{3} - j, -\sqrt{3} - j, 2, j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>00,10</td>
<td>([-\sqrt{3} - j, 2, j, \sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>00,11</td>
<td>([\sqrt{3} - j, -\sqrt{3} - j, 0, 2, j]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>01,00</td>
<td>([2, j, \sqrt{3} - j, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>01,01</td>
<td>([0, 0, -\sqrt{3} - j, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>01,10</td>
<td>([2, j, 0, \sqrt{3} - j, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>01,11</td>
<td>([0, 2, j, -\sqrt{3} - j, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>10,00</td>
<td>([-\sqrt{3} - j, 2, j, 0, \sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>10,01</td>
<td>([\sqrt{3} - j, 0, 2, j, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>10,10</td>
<td>([-\sqrt{3} - j, 0, 0, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>10,11</td>
<td>([\sqrt{3} - j, 2, j, 0, \sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>11,00</td>
<td>([2, j, -\sqrt{3} - j, 0, -\sqrt{3} - j, 0]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>11,01</td>
<td>([0, \sqrt{3} - j, -\sqrt{3} - j, 2, j]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>11,10</td>
<td>([2, j, \sqrt{3} - j, -\sqrt{3} - j, 2]^T/2\sqrt{3})</td>
</tr>
<tr>
<td>11,11</td>
<td>([0, -\sqrt{3} - j, -\sqrt{3} - j, 0, 0]^T/2\sqrt{3})</td>
</tr>
</tbody>
</table>

Since the performance of a SSK-based scheme depends on the minimum distance between any two codewords, an improvement shall be made to the codebook in Table I-(a) to increase its minimum Euclidean distance. To this end, observe that the nonzero elements in Table I-(a), namely \((1 + j)/2\), \(1/2\) and \(j/2\), are resulted directly by multiplexing two in-phase/quadrature GSSK data streams. However, from the perspective of diverting total codeword energy over transmit antennas, these are simply three different complex numbers showing how energies...
are transmitted over different active antennas. Since the minimum distance of the codebook directly depends on the distances among the nonzero elements, the improvement is to replace \((1 + j)/2, 1/2\) and \(j/2\) by three numbers that have the same energy (i.e., magnitude) and maximally spaced in the complex plane, i.e., they lie on the vertices of an equilateral triangle. One set of such numbers are \(-(\sqrt{3} + j), \sqrt{3} - j\), and \(2j\). Let \(n_{RF}\) be the average number of active antennas, which is the ratio between the total number of non-zero elements of all codewords and the total number of codewords. It is simple to show that when the nonzero elements in the codebook have the same magnitude, the normalized minimum squared distance of the proposed codebook is \(d_{\text{min}}^2 = \frac{\sqrt{\alpha}}{n_{RF}}\). For example, for \(N_t = 5\), the total number of non-zero elements of all codewords in the proposed BiGSSK scheme is 202, while the total number of codewords is 64. This gives \(n_{RF} = \frac{202}{64} \approx 3.16\), hence \(d_{\text{min}}^2 = \frac{\sqrt{\alpha}}{3.16}\).

In general, the proposed BiGSSK scheme is constructed as follows:

1. The input bits sequence is divided into two equal-sized groups, \(s_1\) and \(s_2\).
2. Using the GSSK mapping rule with \(n_t = 2\), \(s_1\) and \(s_2\) form two \(N_t \times 1\) symbol vectors \(x_R\) and \(x_I\).
3. Sum \(x_R\) and \(jx_I\) and normalize the sum to form an intermediate transmit symbol vector \(\tilde{x} = (x_R + jx_I)/2\).
4. Change the values \((1 + j)/2, 1/2\) and \(j/2\) in \(\tilde{x}\) to \(-\sqrt{3}/2, \sqrt{3}/2\), and \(2j/2\) to obtain the final transmit symbol vector \(x\), where \(\sqrt{\alpha} = \sqrt{4n_{RF}}\) is the normalizing factor to ensure that \(E(\|x\|^2) = 1\).

As an example, consider a MIMO system with \(N_t = 4\). First, the GSSK mapping rule for activating \(n_t = 2\) antennas out of \(N_t = 4\) available transmit antennas is as follows: \(00 \rightarrow [1, 1, 0, 0]^T, 01 \rightarrow [0, 0, 1, 1]^T, 10 \rightarrow [1, 0, 0, 1]^T, 11 \rightarrow [0, 1, 1, 0]^T\). Now, suppose that a group of four information bits to be transmitted is 1001. First, determine \(s_1 = [1, 0]\) and \(s_2 = [0, 1]\). Using the above GSSK mapping rule, \(s_1\) and \(s_2\) form vectors \(x_R = [1, 0, 0, 1]\) and \(x_I = [0, 0, 1, 1]\). Summing \(x_R\) and \(jx_I\) and normalizing form vector \(\tilde{x} = [1, 0, j, 1 + j]/2\). Changing elements \(1/2\) to \(-\sqrt{3}/2\), \(j/2\) to \(\sqrt{3}/2\), and \((1 + j)/2\) to \(-\sqrt{3}/2\) (here \(\alpha = 12\) and \(n_{RF} = 3\)) gives the transmit symbol vector \(x = [\sqrt{3} - j, 0, 2j, -\sqrt{3} - j]/2\sqrt{3}\). The full symbol mapping for this example is summarized in Table I-(b).

It is pointed out that the GSSK rule used in the proposed BiGSSK can be applied with \(n_t \geq 2\). However, the performance will be degraded with increasing \(n_t\). This present paper focuses on the GSSK with \(n_t = 2\) since this selection of \(n_t\) minimizes the number of RF chains. In general, the spectral efficiency of BiGSSK is \(2\lceil\log_2(n_t)\rceil\) bits/sec/Hz. Table II compares the minimum squared Euclidean

<table>
<thead>
<tr>
<th>(N_t)</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>16</th>
<th>32</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rate</td>
<td>(d_{\text{min}}^2)</td>
<td>Rate</td>
<td>(d_{\text{min}}^2)</td>
<td>Rate</td>
<td>(d_{\text{min}}^2)</td>
<td>Rate</td>
</tr>
<tr>
<td>SSK</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>GSSK ((n_t = 2))</td>
<td>2</td>
<td>1</td>
<td>3</td>
<td>1</td>
<td>3</td>
<td>1</td>
</tr>
<tr>
<td>BiSSK</td>
<td>4</td>
<td>1</td>
<td>4</td>
<td>4</td>
<td>1</td>
<td>4</td>
</tr>
<tr>
<td>BiGSSK</td>
<td>4</td>
<td>(\sqrt{2})</td>
<td>6</td>
<td>(\sqrt{3})</td>
<td>6</td>
<td>(\sqrt{3})</td>
</tr>
</tbody>
</table>
4 Numerical results

For all the BER simulation results versus the average SNR per bit per receive antenna \( (E_b/N_0) \) presented in this section, \( N_r = 3 \) receive antennas are employed. The BERs of BiSSK and BiGSSK with different system settings are shown in Fig. 1-(a). At fixed target transmission rates of 6, 8 and 10 bits/sec/Hz, BiGSSK shows a negligible performance loss (<0.5 dB) but requiring much smaller numbers of available transmit antennas as compared to BiSSK (e.g. for 10 bits/sec/Hz, BiGSSK requires 9 antennas while BiSSK needs 32 antennas). With similar numbers of transmit antennas, the transmission rate of BiGSSK(7,3) is higher, by 2 bits/sec/Hz (which is 33.33% higher), than that of BiSSK(8,3) at the expense of 1 dB performance loss at the BER level of \( 10^{-5} \).

Comparison of LSSK(5,3) and BiGSSK(5,3) for the same number of transmit antennas are shown in Fig. 1-(b). Both LSSK and BiGSSK schemes aim to enhance the transmission rate of SSK under a limited number of transmit antennas. As can be seen, although BiGSSK is inferior to LSSK by less than 0.5 dB, BiGSSK achieves a significantly higher transmission rate (more than 2 bits/sec/Hz, or 50% higher) than LSSK. The performance difference observed in Fig. 1-(b) can be roughly predicted from the minimum squared distances in Table II and the spectral efficiencies of BiGSSK \( (\lambda = 6 \text{ bits/sec/Hz}) \) and LSSK \( (\lambda = 6 \text{ bits/sec/Hz}) \) as follows. Since \( E_s = \lambda E_b \), the power loss of BiGSSK compared to LSSK is \( 10 \log_{10} \frac{6\times10^3}{4\times10} \approx -0.23 \text{ dB} \). This is reasonably close to −0.5 dB observed in Fig. 1-(b).

5 Conclusion

A new SSK-based modulation scheme is proposed by multiplexing in-phase and quadrature GSSK streams and optimizing the signals transmitted over active antennas. Given the same number of transmit antennas, the proposed scheme doubles the transmission rate of GSSK and it achieves a higher transmission rate than BiSSK at a negligible performance loss. Viewed differently, at the same transmission rate, the proposed scheme employs a smaller number of transmit antennas, leading to a reduction in hardware cost.
Millimeter-wave close proximity high-speed data transfer system

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Abstract: This paper presents the system concept, transceiver architecture, and control sequence for a millimeter-wave (60-GHz) band close proximity high-speed data transfer system. The communication range and the use case are limited to achieve fast link setup time and a stable point-to-point connection. Prototype equipment developed for the system includes three types of wireless transceivers; cooperative operation among them makes it possible to reduce the link setup time and limit the communication range. The system’s control sequence enables the link setup time to be reduced from 7 seconds to 0.2 seconds.

Keywords: millimeter wave, 60 GHz, close proximity high-speed data transfer, link setup time, point-to-point connection

Classification: Wireless Communication Technologies

References


1 Introduction

Mobile traffic has increased significantly in recent years with the widespread use of smartphones and tablets and their increasingly high volume content. It is expected that wireless LANs will enable traffic offloading, but in urban areas their effective throughput is rapidly degraded as the user number increases. Millimeter-wave (60-GHz) band wireless systems have a wide bandwidth of about 9 GHz and great...
potential in terms of capacity [1, 2]. There are, however, some problems in using the millimeter-wave band, including link setup time and communication topology. We propose a millimeter-wave close proximity high-speed data transfer system to solve these problems and to provide a comfortable data transmission service. In this paper we describe the system concept, transceiver architecture, and the performance of prototype equipment developed for the system.

2 System concept

The use case of our target system is high-speed data download/upload services that are carried out at kiosk terminals, which are located in many types of public spaces including convenience stores, train stations, and airports. Our target system has the following features:

1) Link setup time is less than 1 second:

Link setup includes association and authentication between a kiosk terminal and a mobile terminal. When the transmission rate of wireless links becomes high, the link setup time will remain noticeably high. For example, a video file with a viewing time of 1 hour that is compressed with a H.265 video compression standard has a data size of about 400 MB. This results in file download/upload time of about 3.2 seconds at a transmission rate of 1 Gbit/s, while the link setup time of the existing millimeter-wave systems is generally more than 5 seconds. Thus the link setup time should be reduced to provide a short file transfer period.

2) The communication range is less than 10 cm:

This short-range system ensures that the system’s communication topology is a point-to-point connection, whereas the existing millimeter-wave systems including IEEE 802.11ad and IEEE 802.15.3c may have point-to-multipoint connections since their communication ranges are several meters [1, 2]. A point-to-point connection has a larger effective throughput than point-to-multipoint connections. The effective throughput of user terminals at an access point or a base station lessens as the number of users per cell increases.

In order to achieve these features, the system we propose combines multiple wireless systems. We hereafter refer to it as the “proposed multiple wireless system” and describe it in the following section.

3 Configuration

The transmitter configuration for our proposed multiple wireless system is shown in Fig. 1. The kiosk terminals and mobile terminals have high-speed caches between which data transfer is carried out. Both the kiosk terminals and mobile terminals have three kinds of wireless transceivers: millimeter-wave, Wi-Fi wireless LAN, and NFC (near field communications).

The transmitter control sequence is shown in Fig. 2. When a user with a mobile terminal enters a Wi-Fi communication zone (about 10 m), Wi-Fi connection and authentication are carried out. The authentication information is used in millimeter-wave connections and NFC connections. When a mobile terminal enters an NFC communication zone (about 10 cm), the terminal is identified. Millimeter-wave data transfer is carried out for the identified mobile terminals.
In this sequence the link setup, including association and authentication between a kiosk terminal and a mobile terminal, will have been finished before a mobile terminal approaching the contact area of a kiosk terminal starts acting. NFC is used to limit the communication range of millimeter-wave data transfer. This is because the NFC antennas operate through electromagnetic induction, which makes it easy to limit the communication range to less than that for antennas that operate through radio wave emission. If the equivalent isotropically radiated power (EIRP) of a millimeter-wave module is set so that the communication range is 10 cm, the millimeter-wave connection within 10 cm may be unstable due to a tilt of a mobile terminal or a position alignment error between a mobile terminal and a kiosk terminal. However, this problem does not occur when the communication range is limited by using NFC.

4 Prototype equipment

We have developed prototype equipment including kiosk terminals and mobile terminals that have millimeter-wave (60 GHz) modules in them as shown in Fig. 3. Between the millimeter-wave modules the transmission rate was 2.5 Gbit/s and between the caches it was 1 Gbit/s, including the read and write times of the caches.
Measured performance comparison between a conventional millimeter-wave module (i.e., one for millimeter waves only) and the proposed multiple wireless system is as follows:

1) Conventional millimeter-wave module:
The link setup time is 7 seconds. The communication range is 50 cm.

2) Proposed multiple wireless system:
The link setup time is 0.2 seconds. The communication range is 10 cm.

As the comparison shows, the latter has much faster link setup time at the designed lower communication range.

5 Conclusion

In this paper, we described the system concept, transceiver architecture, and control sequence for a millimeter-wave (60-GHz) band close proximity high-speed data transfer system. The communication range and the use case are limited to achieve fast link setup time and a stable point-to-point connection. Prototype equipment developed for the system includes three kinds of wireless transceivers, and cooperative operation among them makes it possible to reduce the link setup time and limit the communication range. The system’s control sequence provided 1 Gbit/s transmission rate and enabled the link setup time to be reduced from 7 seconds to 0.2 seconds with the targeted communication range of 10 cm.