A Simulator of Periodically Switching Channels for Power Line Communications

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Abstract. An indoor power line is one of the most attractive media for in-home networks. However, there are many technical problems for achieving in-home power line communication (PLC) with high rate and high reliability. One of such problem is the degradation in the performance of the in-home PLC caused by periodically time-varying channel responses, particularly when connecting the switching power supply equipment. We present a measurement method for power line channel responses and reveal the switching of the channel responses synchronized with power-frequency voltage when connecting switching power supply equipment in sending or receiving outlets. In this paper, we term them periodically switching channel responses. The performance of PLC adapters is seriously affected by the periodically switching channel responses. Therefore, we provide a modeling of the periodically switching channel responses by using finite impulse response (FIR) filters with a shared channel memory and construct a simulator for in-home power line channels including the periodically switching channel responses in order to evaluate the various communication systems through the power line. We present the validity of the proposed simulator through the performance evaluation of OFDM/64QAM over periodically switching channels with additive white Gaussian noise. Furthermore, we evaluate the influence of the periodically switching channel responses on the communication quality of a time-invariant modulation scheme by using the proposed simulator.

Keywords: power line communication, periodically switching channel response, switching power supply, OFDM, simulator.

1 Introduction

An indoor power line is a promising communication medium for constructing in-home networks. In particular, the broadband in-home power line communication (PLC) is receiving increasing attention lately. There are a large number of power outlets for supplying electrical power in any room. Therefore, new wiring for constructing an in-home PLC network is not required, and we can immediately access the PLC network by plugging them into power outlets.

However, users might experience unexpected troubles while communicating with one another in the house by using the indoor power line. Such troubles would be caused

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by the significant signal attenuation through power distribution boards, colored and impulsive noises generated from electrical appliances, the impedance mismatching due to the absence of electrical termination across the frequency band for in-home PLC, and the time-varying channel responses synchronized to twice the electrical power frequency. The technical problems encountered in in-home PLC have been reported in several studies papers: the multipath effect caused by impedance mismatching [1, 2], cyclostationary and impulsive noises [2, 3, 4, 5], and periodically time-varying channel response [6, 7].

Almost all in-home broadband PLC systems make use of the orthogonal frequency division multiplexing (OFDM) since OFDM systems can establish coexistence with the existing radio communication systems via subcarrier masking. Besides, the cyclic prefix of the OFDM could sufficiently mitigate the frequency selective effect caused by the multipath effect since the delay spreads for in-home PLC would not exceeds the length of the cyclic prefix.

Katayama et al. reported the cyclostationary noise that is synchronized to twice the electrical power frequency [3]. Further, Cañete et al. showed the periodically time-varying channel response that is synchronized to twice the electrical power frequency [7], and Sancha et al. presented a channel simulator that could emulate time fluctuations in channel response [6]. Katar et al. asserted that the cyclostationary noise would have a much greater impact on the performance of in-home PLC systems than on the periodically time-varying channel response, and proposed an optimal length of MAC frame and a countermeasure against the beacon loss for cyclo stationary noise in in-home PLC [8, 9].

However, we observed more intensive time fluctuation in the channel responses than in cyclostationary noise when connecting switching power supply equipment [10]. At this time, two or more channel responses switch during one half of a power cycle. In this paper, we term them periodically switching channel responses. These responses severely impair the performance of PLC adapters [10]. Therefore, countermeasures against these periodically switching channel responses are required constructing an in-home PLC network.

In this study, we show a measurement method for power line channel responses and present the periodically switching channel responses when connecting some switching power supply equipment. We provide a modeling of the periodically switching channel responses using finite impulse response (FIR) filters with a shared channel memory and construct a simulator that can emulate the periodically switching channel responses for in-home power line channels. Furthermore, we present the validity of the proposed simulator by evaluating the performance of OFDM/64QAM over the periodically switching channels with additive white Gaussian noise. In addition, by using the proposed simulator we evaluate the influence of the periodically switching channel responses on the communication quality of a time-invariant modulation scheme by using the proposed simulator.

The remaining part of the paper is organized as follows. Section 2 presents a measurement method for power line channel responses and analyzes the measurements by a time-frequency analysis. Section 3 shows a modeling of the periodically switching channel responses and estimates the parameters for the description of a power line channel. Section 4 proposes a simulator for in-home power line channels that can emulate the periodically channel responses and show the validity of the proposed simulator. Section 5 investigates the influence of the periodically switching channel responses to the communication quality by using the proposed simulator. Section 6 presents the conclusion of this study.

2 Measurement of Periodically Switching Channel

This section describes a method for obtaining time-varying channel frequency responses by a time-frequency analysis. Figure 1 illustrates the measurement system for time-varying channels synchronized with a commercial power supply.

In this study, we measure the fundamental power line topology with a single branch. Electrical power is supplied to this power line through a noise cut transformer. A cell-phone charger is connected to the receiving outlet, and thereby, power-line channels become time-varying synchronized with the commercial power supply when the cell phone is charged.

We design a multicarrier signal that has a flat power spectrum density in the frequency range from 2 to 30 MHz and low peak-to-average power ratio (PAPR). This signal is based on the HomePlug AV specification [11], where its sampling frequency F_{s} is 75 MS/s, sampling time $T_{\mathrm{s}} = 1/F_{\mathrm{s}}$ is 13.3 ns, and fast Fourier transform (FFT) size N is 3072. The subcarriers are numbered in order from lowest frequency to highest frequency. The number of subcarriers corresponding to 2 MHz is represented as $k_{\mathrm{f}}(= 82)$, and those corresponding to 30 MHz as $k_{\mathrm{e}}(= 1228)$. The total number of subcarriers $N_{\mathrm{s}} = k_{\mathrm{e}} - k_{\mathrm{f}} + 1$ is 1147. The Z-transform of this signal is given by

$$P(z) = \sum_{i=0}^{N-1} p_i z^{-i}, \tag{1}$$

$$p_i = \mathrm{Re}\left[\sqrt{\frac{2EZ_0}{N_{\mathrm{s}}}} \sum_{k=k_{\mathrm{f}}}^{k_{\mathrm{e}}} \exp\left(\mathrm{j}\left(2\pi\frac{ki}{N} + \theta_k\right)\right)\right], \tag{2}$$

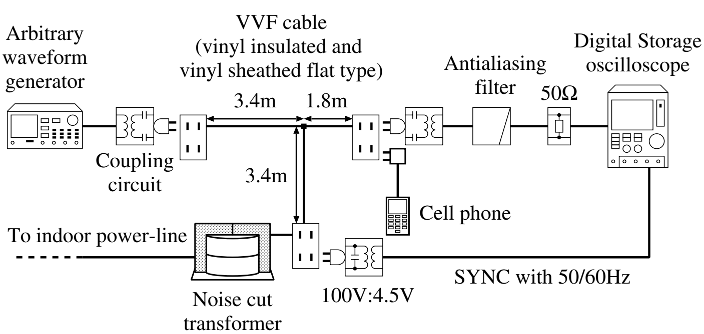

Fig. 1. A measurement system for time-varying channels synchronized with commercial power supply

where E is the energy of this signal, and Z_0 is the impedance of the arbitrary waveform generator (AWG). In order to overcome the effect of quantization noise, we optimize the phase θ_k of each subcarrier to minimize the variance of instantaneous power

$$\frac{1}{N}\sum_{i=0}^{N-1}\left(p_i^2 - \frac{1}{N}\sum_{j=0}^{N-1} p_j^2\right)^2 . \tag{3}$$

The designed signal is continuously transmitted from the AWG to the power line through a coupling circuit. The transmitted signal is received by a digital storage oscilloscope from the power line through the coupling circuit and an anti-aliasing filter. The digital oscilloscope is triggered by an attenuated power-frequency voltage. The coupling circuit comprises a high-pass filter and a balun for eliminating the effect of commercial power supply on the experimental equipment, and the anti-aliasing filter is a low-pass filter for removing aliasing. The digital oscilloscope records the digitized signal comprising a direct wave and some delayed waves because the power line has only one branch and some impedance-mismatched terminals.

The Z-transform of the transmitted signal is given by

$$S(z) = \sum_{\ell=0}^{\infty}\sum_{i=0}^{N-1} p_i z^{-i-N\ell} = \sum_{\ell=0}^{\infty} P(z) z^{-N\ell}, \tag{4}$$

and the Z-transform of the received signal is given by

$$R(z) = \sum_{\ell=0}^{\infty} H_\ell(z)P(z)z^{-N\ell} + N(z), \tag{5}$$

$$H_\ell(z) = \sum_{j=0}^{D-1} h_{\ell,j} z^{-j}, \tag{6}$$

where $H_\ell(z)$ represents the impulse response for the ℓ-th signal $P(z)$, $N(z)$ indicates the additive power line noise, and DT_s represents the maximum multipath delay (we assume $D < N$). The noise $N(z)$ can be approximated to zero because we exploit the noise cut transformer. The ℓ-th received signal $R_\ell(z)$ is given by

$$R_\ell(z) = \sum_{m=N}^{N+D-2}\ \sum_{j=m-N+1}^{D-1} h_{\ell-1,j}p_{m-j}z^{N-m} + \sum_{m=0}^{N-1}\sum_{j=0}^{D-1} h_{\ell,j}p_{m-j}z^{-m}. \tag{7}$$

Further, when $H_\ell(z)$ is equal to $H_{\ell-1}(z)$, $R_\ell(z)$ is expressed as

$$\begin{aligned} R_\ell(z) &= \sum_{m=0}^{D-2}\sum_{j=m+1}^{D-1} h_{\ell,j}p_{m+N-j}z^{-m} + \sum_{m=0}^{N-1}\sum_{j=0}^{D-1} h_{\ell,j}p_{m-j}z^{-m} \\ &= \sum_{m=0}^{N+D-2}\sum_{j=0}^{D-1} h_{\ell,j}p_{m-j}z^{-m} = H_\ell(z)P(z), \end{aligned} \tag{8}$$

where $z^{-m} \equiv z^{-m-iN}$ for any integers m and i. Therefore, the frequency response is given by

$$H(t,f)|_{t=\ell T} = H_\ell\left(\exp\left(\mathrm{j}\frac{2\pi f}{F_\mathrm{s}}\right)\right) = \frac{R_\ell\left(\exp\left(\mathrm{j}\frac{2\pi f}{F_\mathrm{s}}\right)\right)}{P\left(\exp\left(\mathrm{j}\frac{2\pi f}{F_\mathrm{s}}\right)\right)}, \tag{9}$$

where T is the duration of the designed multicarrier signal, and it is equal to NT_s.

Figure 2 illustrates the time-frequency analysis of the power line channel during two power cycles. The origin at the horizontal axis indicates an increasing in the

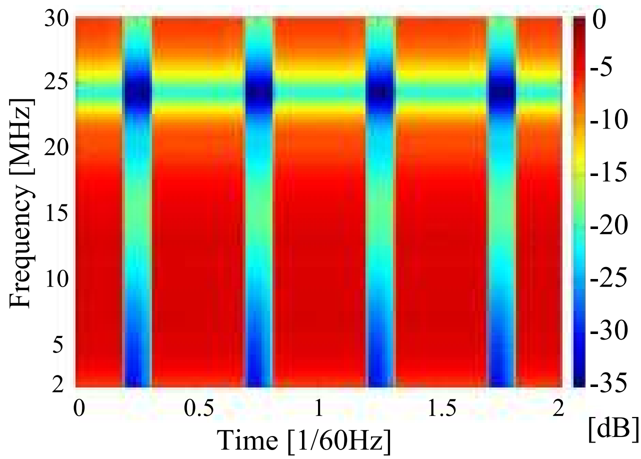

Fig. 2. Periodically switching channel responses

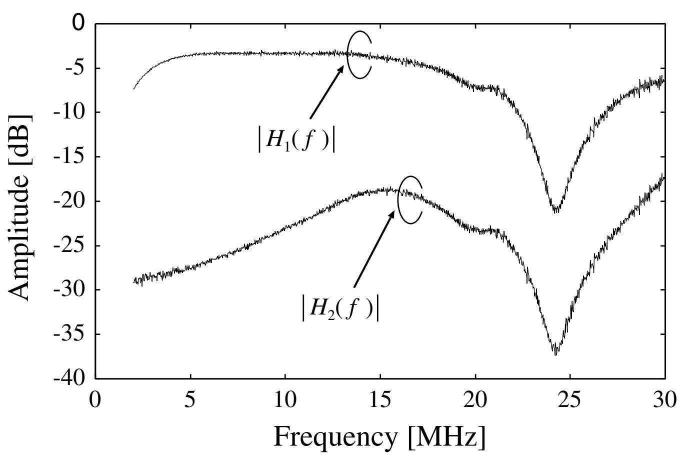

Fig. 3. Amplitude spectra of two channel responses

zero-crossover point of the commercial power supply. This figure shows that while charging a cell phone, the power line channel could be classified into almost two states and the duration of the state with higher loss is related to the timing when the instantaneous voltage of power supply comes near the peak. Let us indicate the state with lower loss as state 1 and that with higher loss as state 2, and their corresponding frequency responses as $H_1(f)$ and $H_2(f)$, respectively. Figure 3 illustrates the amplitude spectra of the channel responses of the two states. Although both the channel responses seem to be correlated, they are actually different, and hence, modeling of the responses required.

3 Modeling of Periodically Switching Channel

Indoor power line channels will be connected to several electrical appliances. Hence, the PLC signals reflect on the branches and terminals. As a result, the power line channels become multipath. The modeling of static channel responses for the power line channels is based on the following multipath model [1]

$$H(f) = \sum_{i=0}^{L-1} g_i \mathrm{e}^{-(a_0 + a_1 f^k) d_i} \mathrm{e}^{-\mathrm{j}2\pi f(\tau_i - \tau_0)}, \tag{10}$$

with

$$\tau_i = \frac{d_i \sqrt{\varepsilon_{\mathrm{r}}}}{\mathrm{c}} = \frac{d_i}{v_{\mathrm{p}}}, \tag{11}$$

where the parameters in the expressions are listed in Table 1.

Table 1. Parameters of the frequency response

L	total number of the dominant paths
g_i	weighting factor for i-th path
a_0	attenuation constant at constant term (> 0)
a_1	attenuation constant at term of f^k (> 0)
k	exponent of the attenuation factor (between 0.2 and 1)
τ_i	delay of i-th path ($\tau_0 < \tau_1 < \cdots < \tau_{L-1}$)
d_i	length of i-th path
ε_{r}	dielectric constant
c	speed of light in vacuum ($3.0 \times 10^8 \mathrm{m/s}$)
v_{p}	phase velocity

However, this model does not include the short-term variation of the channel responses. Considering the variation caused by the two channel states, the frequency response can be represented as

$$H(t, f) = \begin{cases} H_1(f), & \text{if } |v_{\mathrm{AC}}(t)| \leq V_{\mathrm{th}}, \\ H_2(f), & \text{if } |v_{\mathrm{AC}}(t)| > V_{\mathrm{th}}, \end{cases} \tag{12}$$

for power supply voltage $v_{\mathrm{AC}}(t)$, where V_{th} represents the threshold of instantaneous voltage switching from one state to another state.

We assume that the frequency responses $H_1(f)$ and $H_2(f)$ can be represented by the above mentioned model. For $m = 0$ and 1, let us represent $H_m(f)$ as

$$H_m(f) = \sum_{i=0}^{L-1} g_i^{(m)} \mathrm{e}^{-(a_0^{(m)} + a_1^{(m)} f^{k^{(m)}})} \mathrm{e}^{-\mathrm{j}2\pi f(\tau_i^{(m)} - \tau_0^{(m)})} \ . \tag{13}$$

Let $h_m(t)$ represent the impulse response of the frequency response $H_m(f)$, and $h_j^{(m)} = h_m(jT_{\mathrm{s}})$ for $j = 0, 1, \cdots, D-1$. The Z-transform of $h_m(t)$ is denoted by $H_m(z)$, and we can obtain

$$H_m(z)\Big|_{z=\exp\left(\mathrm{j}\frac{2\pi f}{F_{\mathrm{s}}}\right)} = H_m(f). \tag{14}$$

We estimate the model parameters of $H_m(z)$, which are shown in Table 1, from the measurement data for a given number of paths L by a quasi-Newton method. Let us consider the measurement value $H_m(kF_{\mathrm{s}}/N)$ at frequency kF_{s}/N for $k_{\mathrm{f}} \leq k < k_{\mathrm{e}}$. We can obtain the estimated frequency response $\hat{H}_m(f)$ to minimize the mean square error (MSE)

$$\sum_{k=k_{\mathrm{f}}}^{k_{\mathrm{e}}} \left| \hat{H}_m\left(\frac{kF_{\mathrm{s}}}{N}\right) - H_m\left(\frac{kF_{\mathrm{s}}}{N}\right) \right|^2 \ . \tag{15}$$

We are required to select an approximate number of paths L. If L is large, the estimated frequency response will sufficiently fit the exact frequency response, however, a long computational time is consumed. On the other hand, if L is small, the approximation of the estimated frequency response to the exact frequency response will lack precision. Figure 4 illustrates the MSE between the estimated and measured frequency responses for a given number of paths.

From Fig. 4, we assign the number of paths as 8. Let $\hat{H}_m(f)$ be the frequency response obtained by fitting the model parameters to the measured frequency response $H_m(f)$, and $\hat{H}(t, f)$ be the estimated periodically switching channel responses consisting of $\hat{H}_1(f)$,$\hat{H}_2(f)$, and V_{th}.

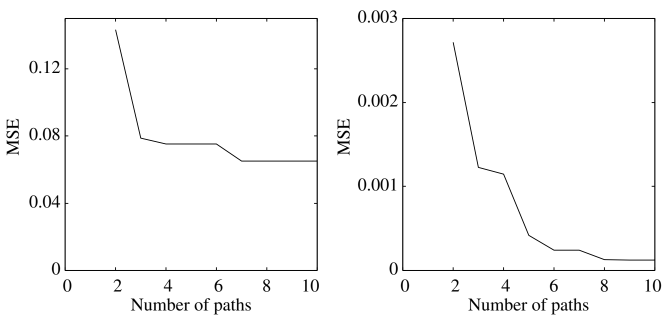

Fig. 4. MSE of the estimated frequency response for a given number of paths

4 Channel Simulator

It is effective to construct a channel simulator that emulates the power line channel for developing the PLC systems. Therefore, we propose a simulator using the parameters estimated in Section 3 and two FIR filters with shared channel memory. The two FIR filters are switched according to V_{th}. Figure 5 shows the block diagram of the simulator for periodically switching channels.

To evaluate the validity of this simulator, we first compare the amplitude and the phase of $\hat{H}_m(f)$ of the simulator with those of $H_m(f)$.

Figures 6 and 7 illustrate the amplitude and phase of $H_m(f)$ and $\hat{H}_m(f)$, and Table 2 shows the model parameters for $\hat{H}_1(f)$ and $\hat{H}_2(f)$. These figures show that the amplitude of $\hat{H}_m(f)$ is sufficiently approximated to that of $H_m(f)$, whereas the phase

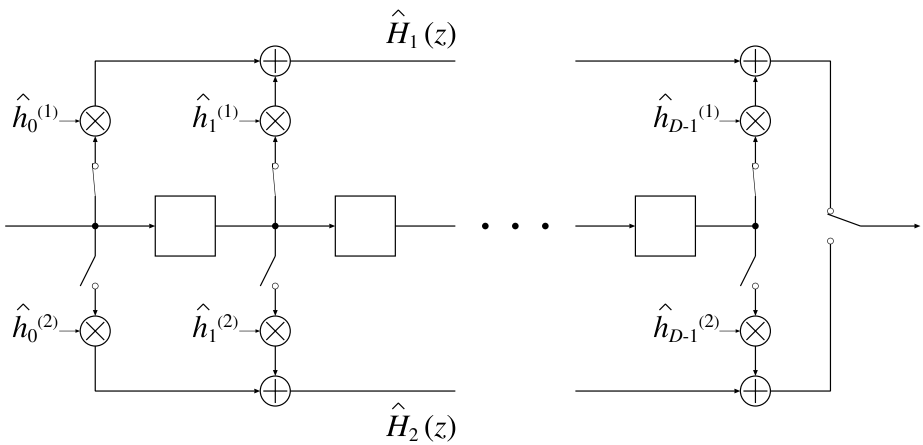

Fig. 5. The block diagram of the simulator for periodically switching channels

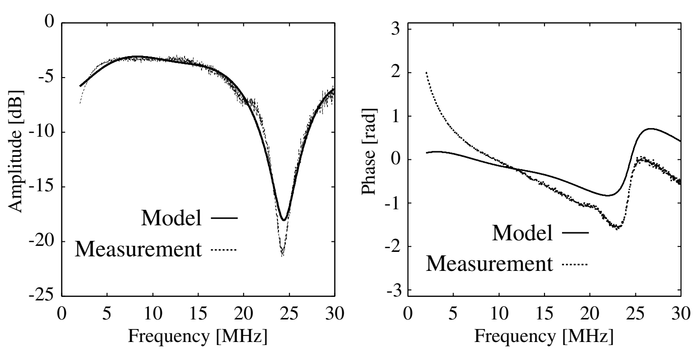

Fig. 6. The spectra of the estimated frequency response $\hat{H}_1(f)$ and the measured frequency response $H_1(f)$

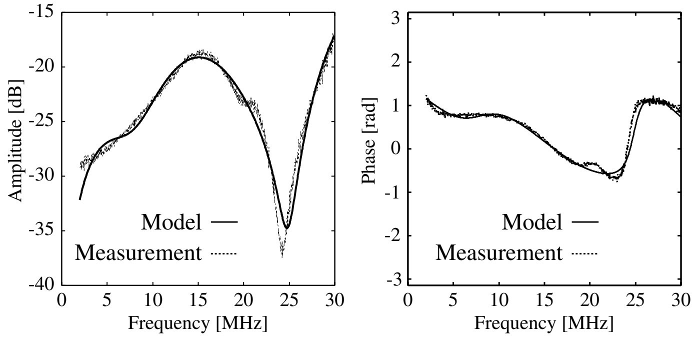

Fig. 7. The spectra of the estimated frequency response $\hat{H}_2(f)$ and the measured frequency response $H_2(f)$

Table 2. Model parameters for the estimated frequency responses $\hat{H}_1(f)$ and $\hat{H}_2(f)$

	$\hat{H}_1(f)$	$\hat{H}_2(f)$
L	8	8
g_i	$1.00, 0.277, -0.0712, -1.00$ $-0.0792, 0.561, -0.784, 0$	$0.0660, -0.0341, 0.266, -0.328$ $0.434, -0.401, 0.156, -0.123$
a_0	4.33×10^{-13}	5.20×10^{-11}
a_1	4.60×10^{-3}	1.59×10^{-7}
k	0.205	0.599
d_i	$5.20, 8.80, 12.0, 12.4, 15.6, 16.0, 18.8, 19.2$	
v_p	1.73×10^{8}	1.73×10^{8}

of $\hat{H}_m(f)$ may not be sufficiently approximated to that of $H_m(f)$. Therefore it is necessary to consider the effect on the communication system by using the difference between the phase of $\hat{H}_m(f)$ and $H_m(f)$.

We evaluate the simulator by using a typical modulation scheme for the in-home PLC. We utilize the OFDM and the quadrature amplitude modulation (QAM) as a subcarrier modulation, which is adopted in HomePlug AV [11] and UPA [12]. Figure 8 illustrates the block diagram of the in-home PLC system, and Table 3 lists the simulation parameters.

Table 3. Simulation parameters

sampling frequency	75 MHz
number of FFT points	3072
number of subcarriers	1147 (2–30 MHz)
number of cyclic prefix	417
subcarrier modulation	4, 16, 64QAM
average noise power	-42.6 dBm

The transfer function shown in Fig. 8 indicates the measured periodically switching channel response $H(t, f)$ or the estimated periodically switching channel response $\hat{H}(t, f)$. We compare the bit error rate (BER) of the measured channel response $H(t, f)$ with that of the estimated channel response $\hat{H}(t, f)$. In the simulation, a random binary sequence is mapped to the QAM symbol sequence. The QAM symbol sequence is modulated by OFDM. The resultant OFDM/QAM signal is inputted to the periodically switching channels. The change the frequency response is obtained by switching the two frequency responses when measuring the frequency response changes. The white Gaussian noise is added to the signal outputted from the channel, and the received signal is equalized with respect to each subcarrier and demodulated. In general, the evaluation of modulation schemes would be done by BER versus received signal-to-noise ratio (SNR). However, BER versus received SNR is not suitable for the evaluation of

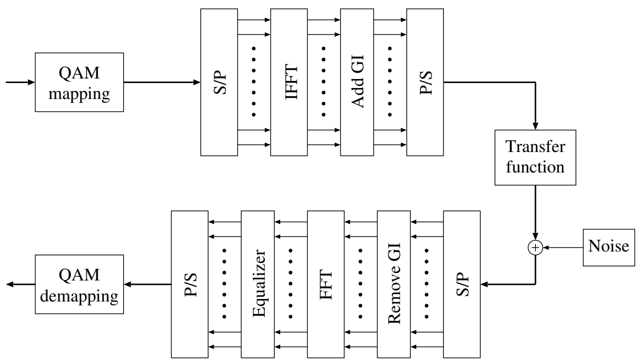

Fig. 8. The block diagram of the in-home PLC system with the proposed channel simulator

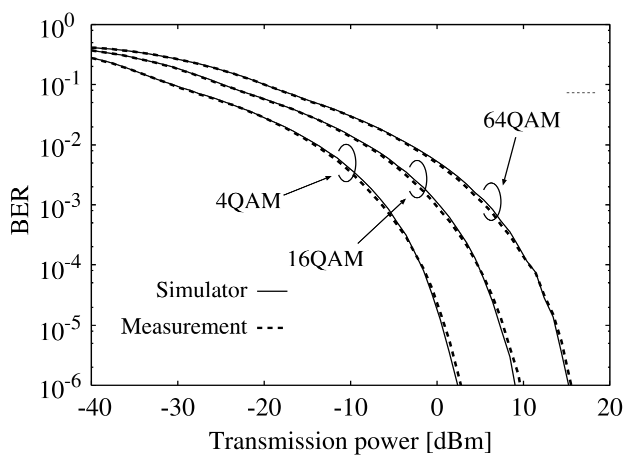

Fig. 9. BER performances for the measured channel response $H(t, f)$ and the estimated channel response $\hat{H}(t, f)$ by using the proposed channel simulator

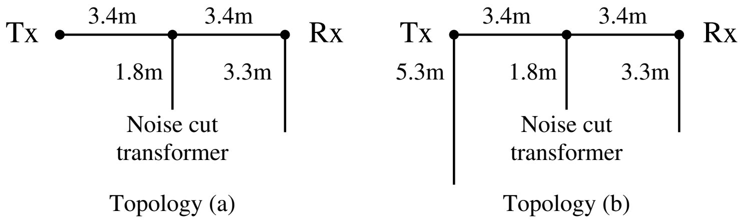

Fig. 10. Topologies

different channel responses. Hence, in this study, we add the white Gaussian noise, which includes the noise power obtained by measuring the power line noise, and evaluate the BER versus the transmission power. We measure the noise on the power line connected some desktop computers in our laboratory. For a bandwidth of 28 MHz, the average noise power measured is -42.6 dBm, and the one-side power spectral density N_0 is -117.0 dBm/Hz. Figure 9 illustrates the comparison between the BER and the transmission power using the measured and estimated channel responses.

This figure shows that the BER for the estimated channel response is almost equal to that for the measured channel responses regardless of the phase difference, as can be seen in Fig. 6. This is because the OFDM with guard interval mitigates the phase difference. Therefore, it would be appropriate to simulate OFDM communication systems with the power line channels with periodically varying frequency responses by using the proposed channel simulator.

Further, we apply two additional topologies (a) and (b), as shown in Fig. 10, to our proposed channel simulators. Figures 11 and 12 illustrate the spectra of the estimated frequency response $\hat{H}_1(f)$ and the measured frequency response $H_1(f)$ of topology (a). As well, Fig. 13 and 14 illustrate the spectra of the estimated frequency response $\hat{H}_1(f)$ and the measured frequency response $H_1(f)$ of topology (b). The estimated spectra of topologies (a) and (b) match excellently with the measured spectra of topologies (a) and (b), respectively. Therefore, we conclude that our proposed simulators could describe a lot of actual power line channels.

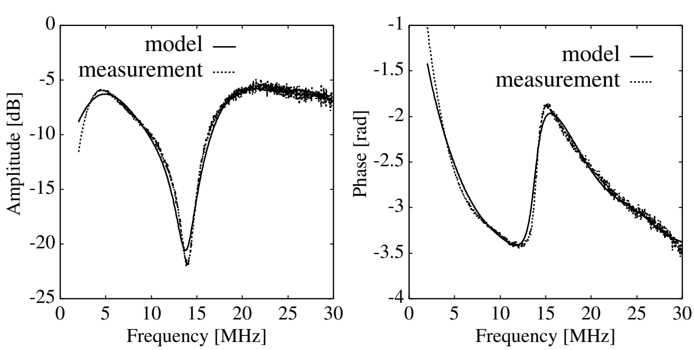

Fig. 11. The spectra of the estimated frequency response $\hat{H}_1(f)$ and the measured frequency response $H_1(f)$ of topology (a)

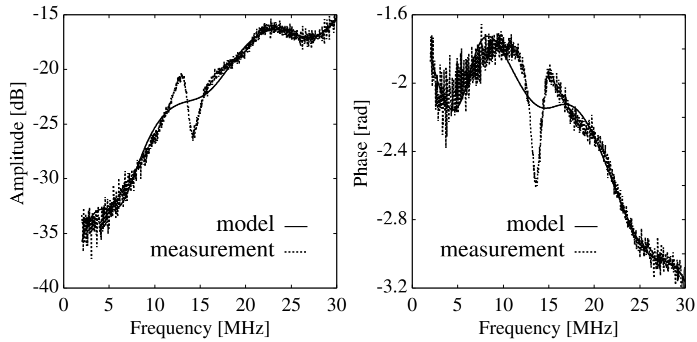

Fig. 12. The spectra of the estimated frequency response $\hat{H}_2(f)$ and the measured frequency response $H_2(f)$ of topology (a)

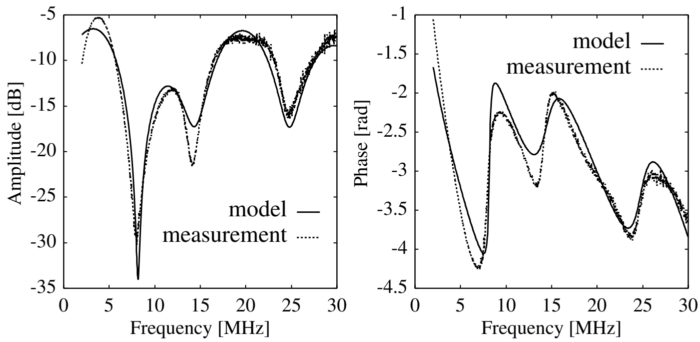

Fig. 13. The spectra of the estimated frequency response $\hat{H}_1(f)$ and the measured frequency response $H_1(f)$ of topology (b)

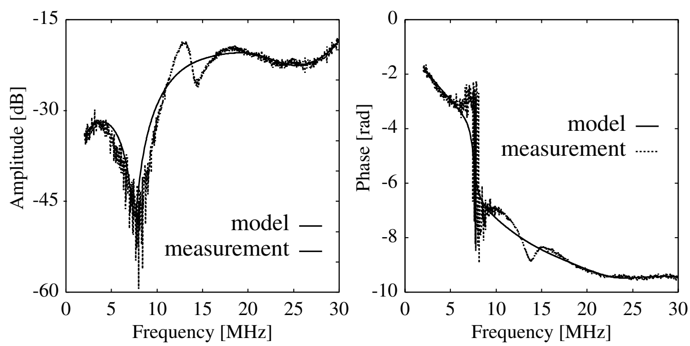

Fig. 14. The spectra of the estimated frequency response $\hat{H}_2(f)$ and the measured frequency response $H_2(f)$ of topology (b)

5 Effect of Incorrect Channel Estimation

There might exist short-term variations in the channel response of in-home PLCs. In the case where the channel is estimated to be time-invariant but the actual channel is a periodically switching channel synchronized with the power frequency voltage, equalization in the receiver would be erroneous, resulting in a drastic increase in the BER. By using the channel simulator (Section 4), we can examine the influence of the channel responses on the communication quality. We evaluate the BER performance when equalized by $\hat{H}_1(f)$ in the entire interval, when equalized by $\hat{H}_2(f)$ in the entire interval, and when equalized by $\hat{H}_1(f)$ and $\hat{H}_2(f)$ in each state correctly. We perform the simulation in these cases and show a comparison between the BER and the transmission power in Fig. 15.

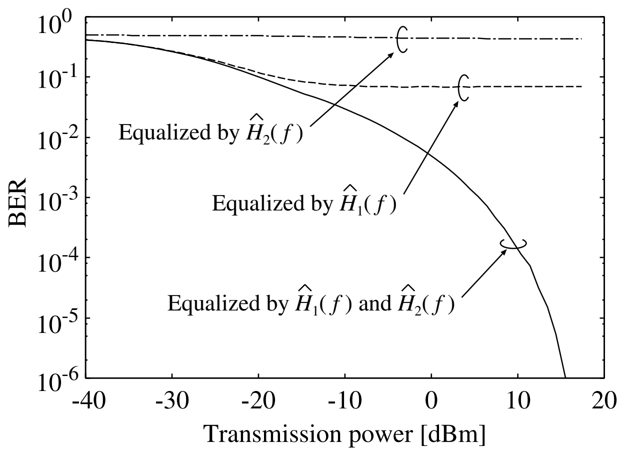

Fig. 15. Evaluating time-varying channel

When the periodically switching channel is estimated to be time-invariant, bit errors would randomly occur for an incorrect equalization. Assuming that the BER can be approximated to zero for a correct equalization, we can obtain

$$\mathrm{BER}_{\mathrm{eq1}} = 0.5 \times \frac{T_2}{T_1 + T_2} = 6.63 \times 10^{-2}, \tag{16}$$

$$\mathrm{BER}_{\mathrm{eq2}} = 0.5 \times \frac{T_1}{T_1 + T_2} = 4.34 \times 10^{-1}, \tag{17}$$

where T_1 and T_2 denote the time interval of state 1 and state 2, respectively. $\mathrm{BER}_{\mathrm{eq1}}$ denotes the BER when equalized by $\hat{H}_1(f)$ over the entire interval, and $\mathrm{BER}_{\mathrm{eq2}}$ denotes the BER when equalized by $\hat{H}_1(f)$ over the entire interval. Figure 15 shows that the BER can be approximated to these values as the transmission power is large. Hence, in the case where the periodically switching channel is estimated as a time-invariant channel, we can roughly estimate the BER if the time intervals of both the two states are known. The simulation results show that the BER in the periodically switching channels increases if the time-varying channel response is not estimated in every state and the equalization would be incorrect.

6 Conclusion

In this study, we showed a measurement method of periodically switching channels that are observed in power line channels. By using the frequency responses obtained by the analysis, we proposed a modeling of the periodically switching channels. We constructed a simulation system for examining the effect of periodically switching channels and evaluated the validity of the proposed simulator. Furthermore, we showed the effect of estimating periodically switching channels as a time-invariant channel. As a result, we could simply evaluate the performance of time-invariant equalization schemes by using the switching intervals of the periodically switching channels.

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