Baseband Transceiver Design of a 128-Kbps Power-Line Modem for Household Applications

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***Abstract*—Communication using power line as a medium provides a convenient and inexpensive way for data transmission and control signaling in the households. In this paper, we will introduce a power-line channel model as well as architecture of a spread-spectrum baseband transceiver integrated circuit (IC) for a power-line modem. The modulation and spreading scheme used in the proposed transceiver is M-ary bi-orthogonal keying (MBOK). This transceiver runs at a chipping rate of 256 kHz and provides 128-Kbps data rate. Simulation results as well as FPGA emulation verify the effectiveness of the proposed architecture for household data communication.**

***Index Terms*—Home automation, modem, power line, spread spectrum.**

I. Introduction

Electric power line, which can be found in most buildings, naturally exhibits its potential as a convenient and cheap communication medium. A reliable intrabuilding communication network based on power line will serve as the most suitable infrastructure for home and office automation. Although some power-line communication systems have already been built and tested, their data rate are usually low and/or their complexity too high for home automation applications [1]–[3].

A. Applications of Power-Line Communication

There are many household applications that power-line communication can serve. These applications fall into two categories: control signaling and data transmission. Control signaling involves low data rate communication between transmitters and receivers, and most home automation applications belong to this category. A typical household scenario of home automation will have all the electrical appliances connected to the power line; control signals transmitted on the power line can reach all of them. A person in the bedroom can thus turn on the light in the living room, reset the air conditioner, or switch off the washing machine easily on a control panel without leaving his room. In addition, temperature sensors can monitor the temperature everywhere in the house and report via the power line to a central controller and, through the power line, the controller can dynamically adjust the air conditioning or trigger a fire alarm if an unexpectedly high temperature is reported.

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The existing power line in a house can also be used as a local area network (LAN) backbone. Computers, modems, and telephones can thus communicate with one another through this network. These data transmission applications and network operations essentially require reliable and high-speed communication capability. Though power-line channel does not offer perfect communication quality for large and sophisticated network operations, it provides a sufficient and low-cost solution to household data communication applications.

B. Residential Power Line

Every communication medium has its own features. Usually they differ in capacity, range, availability, cost, and communication quality. Convenience and low costs are the two most attractive features for residential power line as a communication medium. Since power lines are setup in almost every building, where communication is in great demand, and outlets are distributed evenly, transmitters and receivers with power-line communication ability can, wherever they are and without dedicated cable installation, gain access to the network by simply plugging into nearby power outlets. When an electrical appliance, once used in the living room, is moved to the bedroom, it is still connected to the network as soon as it is plugged in. If a power-line communication channel does not exist, a relocation of a transceiver will involve not only moving the transceiver itself but also rewiring of dedicated communication medium, like twisted-pair or coaxial cables. Sometimes this costs a lot of time and money besides the inconvenience caused by disconnection of the transceiver from the network.

However, because it is designed for power delivery rather than for signal transmission, the power line has several less-than-ideal properties as a communication medium. These include time- and frequency-varying attenuation, 60-Hz power-wave dependent line impedance, and sudden change of channel characteristics. Besides these disadvantages, there are several types of noises, like continuous-wave noise, impulse noise, and transient disruption caused by electrical loads. The background noise present on the power-line channel is also very large, and its level depends on frequency.

Although these distortion and noises may be present also on other communication media, various loads connected onto and disconnected from the power line randomly make the fast changing channel condition even more unpredictable. In addition, the SNR at the receiver end can be very low because the transmitter is far away while a large noise source may be nearby. Even though designing a communication system over this hostile channel is possible, its hardware may be very complicated and expensive and the communication protocol too sophisticated to implement. All these diminish the advantages of the power-line channel and have thus deterred people from making full use of it. Recently, armed with modern micro-electronics and communication technologies, researchers are able to develop feasible and efficient methods to circumvent these obstacles and to take full advantage of its ubiquity.

Currently there are two standards regulating the frequency band and power level of communication signals on power-line communication. One of them was drawn up by the European community standards setting body for electric utilities (CENELEC). The CENELEC EN 50065 standard [4] allows communication over the low-voltage distribution power line between electrical utilities with signal frequency from 3 kHz up to 95 kHz. Compared with the strict European standard, the other power-line communication standard, established by the United States communication regulatory body, the Federal Communications Commission (FCC), mandates more relaxed regulations in the usable frequency band and transmission power [3]. Moreover, this standard is also adopted in Japan [5]. The frequency band allowed by the FCC standard for communication signal ranges from 10 kHz to 450 kHz.

In this paper, we intend to develop a power-line communication modem for both control signaling and data transmission in households. The modem should provide at least 128 Kbps data rate with a BER less than 10^{-4} in the frequency band from 10 kHz up to 450 kHz as specified by the FCC.

II. Power Line Channel Model

Originally designed for power delivery instead of signal transmission, power line possesses impairments as a communication channel, including time- and frequency-varying attenuation and sudden change of channel characteristics. The standing wave effect and multipath interference depending on the wiring of the power line and electric loads should also be considered. In addition, various inimical noises, such as impulse noise and continuous-wave noise produced by electric loads, and large background noise, are also present in power lines.

After extensive literature survey, we conclude that there is no universally-recognized power-line channel model available and the measurement data of the power-line channel comes from various sources [5]–[13]. The measured results and their interpretation may not be consistent and sometimes are contradictory. For instance, some say that the attenuation at a certain frequency can be as large as 40 dB while others report that attenuation more than 30 dB is never observed. Some may conclude that PSK modulation can be used because the phase distortion on the channel is small, while others decide to use FSK for the phase jitter of received signal can be very large. In this paper, we draw several conclusions from these findings and then build a power-line channel model.

Taking all the aforementioned effects into account, we propose a power-line channel model for system simulation and performance evaluation. Its block diagram is shown in Fig. 1. In this model, the time-varying channel filter accounts for frequency-dependent and time-varying attenuation and 120-Hz impedance modulation. The noise sources generate AWGN, impulse noise, and continuous-wave noise to be inflicted upon the filtered signal. The control unit dynamically adjusts several parameters influenced by the 60-Hz power wave. Finally, the receiver prefilter represents a simplified and idealized receiver front-end that filters the signal in the frequency band of interest and pass it to the baseband receiver. In the following, we will explain each block in detail.

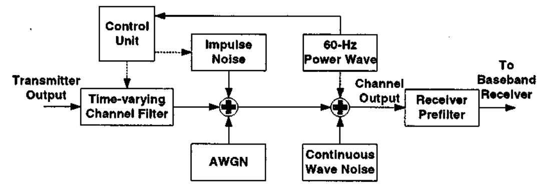

Fig. 1. Block diagram of the power line channel model.

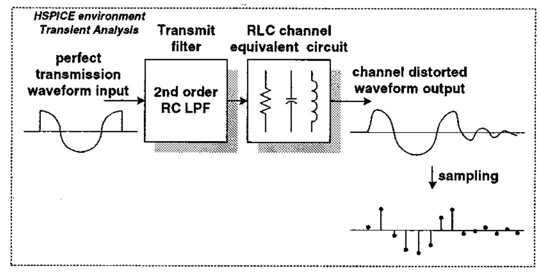

Fig. 2. Derivation of channel filter impulse response.

A. Time-Varying Channel Filter

The channel filter receives baseband transmission signal and outputs the corresponding channel-filtered passband signals. The channel-filtered signals are derived from SPICE transient analysis by sampling the responses of a power-line equivalent circuit with perfect transmission signal as input. The power-line equivalent circuit is constructed using a second order RC low-pass transmit filter, RLC models of some electric apparatus introduced in [13], and models of power-line wiring. The channel filter accounts for the effects of signal attenuation, distortion, and fading in the channel.

The channel filter receives baseband transmission signals and outputs corresponding channel-filtered passband signals. The channel-filtered signal samples are derived through simulation in SPICE and Fig. 2 depicts this procedure. First a perfect passband QPSK symbol passes through a second-order RC low-pass transmit filter with 3-dB frequency at 450 kHz, and the output then enters a power-line equivalent circuit consisting of linear lumped RLC elements. By transient analysis using SPICE, the responses of the combination of transmit filter and power-line equivalent circuit can be obtained, and the response represent the channel-filtered and distorted signal received by the receiver. The waveform of the equivalent circuit response is sampled and stored as an impulse response of the channel filter.

The second-order low-pass filter represents the transmit filter in the transmitter. It is used to attenuate signals beyond 450 kHz and limit the out-of-band transmission power. In conventional narrow-band digital communication systems, a digital raised-cosine filter is utilized to produce band-limited signals with controlled intersymbol interference (ISI). In the proposed system, which is less sensitive to ISI, a second-order low-pass filter suffices.

The power-line equivalent circuit is constructed using the equivalent networks of various kinds of loads, cables, and distribution transformers found in [5] and [6], thus they can correctly simulate the property of power line. These equivalent RLC networks are coded into SPICE netlists, and then connected to each other to form a complete power-line equivalent circuit in accordance with common wiring structures in a household. Thus, in changing the types of loads and altering the configuration of interconnections, many equivalent circuits corresponding to different channel conditions can be easily obtained.

Many impulse responses corresponding to different equivalent circuits are sampled and stored. The channel filter model will dynamically change its coefficients among these sets of responses to produce time-varying effect of the channel. The time-varying property of the power-line channel results from changing loading profile. When a load connects to or disconnects from the power line, the power-line impedance will exhibit a drastic and abrupt change; on the other hand, if no loads are switched on or off in a period of time, the power-line impedance remains stable. Because there are usually dozens or even hundreds of electric appliances in households sharing the same distribution transformer, the number of loads switching on or off a power line in a time duration is random and can be approximated by a Poisson random variable. In the power-line channel model, the minimum unit of time is a sample duration, which is on the order of one microsecond, so the average number of load switches in such a short period of time is quite small. Hence, in a sample duration, the abrupt line impedance change can be determined by the outcome of a Bernoulli trial with a success probability equal to average number of load changes in a sample duration.

B. White Noise, Impulse Noise, Continuous-Wave Noise

The AWGN generator consists of a Gaussian random number generator and some control and arithmetic operations. It can be configured to produce AWGN signals with any mean and noise spectral density.

The noise impulses are the transient spikes generated when load switched on or off the power line. Similar to the sudden change of power line impedance, their occurrence in a time duration can be approximated by a Poisson random variable. The generation of impulse noise in the channel model is quite straightforward. Outcome of a Bernoulli trial with a success probability equal to the average number of impulses in a sample period determines whether an impulse will occur in this sample duration. If so, its polarity, duration, and amplitude are determined by the other three Gaussian random number generators. The shape of the generated impulse noise is triangular and the occurrence probabilities and distribution parameters for polarity, duration, and amplitude can be adjusted.

There are several sinusoidal and triangular waveform generators in the continuous wave generator. They can be configured to produce CW signals with given amplitudes and phases at desired frequencies. The outputs of these waveform generators are summed together to form continuous wave interference signal. Fig. 3 shows a sample output spectrum of the continuous wave generator.

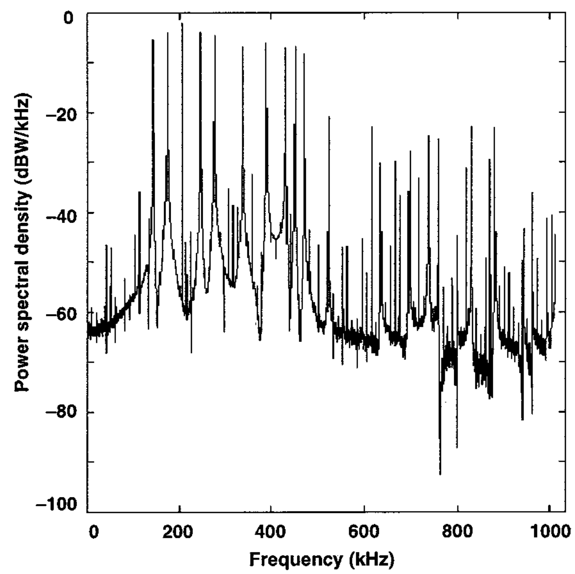

Fig. 3. Output spectrum of continuous wave generator.

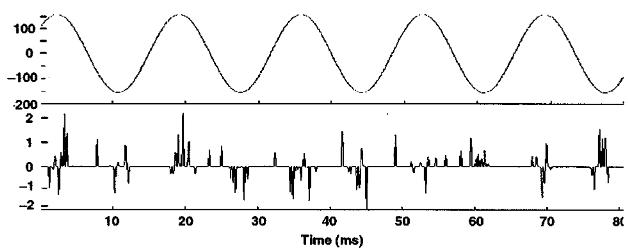

Fig. 4. Power wave and impulse noise output.

C. 60-Hz Power Wave and Control Unit

The dominant 60-Hz power wave, with its relatively low frequency, is in itself harmless to power-line communications, its phase, however, has a great impact upon signal transmissions. As previously described, many loads periodically switch on and off the power line when the power wave is near its peaks. Hence, the effective impedance and attenuation also change at that time, and this periodical change of channel characteristics is called 120-Hz impedance modulation. In addition, frequent connections and disconnections of loads increase the occurrence probability of impulse noises during peaks of the 60-Hz wave.

The control unit implements both effects. It continuously monitors the voltage of the 60-Hz power wave. When the power voltage is near its peaks, the control unit dynamically increase both the probability of impulse response switching in the channel filter and the occurrence probability of impulse noises. When the power wave leaves it peaks, the control unit then decreases these probabilities (see Fig. 4).

Finally, a digital FIR filter representing the band-pass filter in the analog front end of the receiver is also included in the channel model.

D. Channel Conditions

During simulation, the parameters of this channel model, e.g., power spectral density of AWGN, power of continuous-wave noises, and occurrence probability of impulse noises, are set according to the measured values in [5], [10], [12], and [13].

All the noises and impairments introduced previously are present simultaneously on the power-line channel, and they have different impacts upon the communication performance. For convenience in simulation and evaluation of the performance of the proposed power-line modem, we categorize the channel condition into four scenarios, according to the following parameters:

1) frequency of abrupt changes in line impedance;
2) occurrence probability of impulse noises;
3) variance of impulse noise amplitude;
4) power level of continuous-wave noise.

The background noise level, i.e., noise spectral density of AWGN, is used as a variable during simulation so that bit error rate or other performance indices versus the AWGN noise level can be obtained for each of the four possible channel scenarios.

The four channel scenarios include

1) AWGN-only channel;
2) best-case channel;
3) typical channel;
4) worst-case channel.

The AWGN-only channel, with a flat frequency response and no noises except for the AWGN, can be used for preliminary functional simulation and the performance simulation results also serve as a reference to evaluate system performance under other three channel scenarios. The best-case channel stands for a well-behaved power-line channel that can be found in either a simple household environment with only a small number of electric appliances and plain wiring, or ordinary residential buildings when most electric loads are turned off. In this situation, we expect that the continuous-wave noise is small, impulse noises are rare and have small amplitudes, channel impedance may remain stable for a longer time, and attenuation is also insignificant.

The typical channel condition represents a commonly experienced residential power-line channel environment. It accounts for most power-line channel found in a residential building during daytime and evening, and with no severe-noise-producing loads such as drills, light dimmers, or vacuum cleaners present on the power line. Because many electrical appliances are active, recurrent change of channel properties and moderate attenuation are observed, and a considerably high level of noise power is injected into the channel. However, when the attenuation is serious and channel becomes fast-changing, or continuous wave noise and impulse noise are large and occur very frequently due to insertion of a large number of electric loads, the power-line channel becomes incapable of reliable communication. The worst-case channel scenario represents the power-line channel under this situation.

III. Baseband Transceiver Architecture

As shown in Fig. 5, a power-line communication system basically consists of a transmitter, the power-line channel, and a receiver. In this section, we focus on the design and implementation of the baseband transmitter and receiver.

A. Modulation Scheme and Transmitter Architecture

Modern spread-spectrum techniques are adopted to combat the hostile power-line channel. After examining performances, spectral efficiency, and hardware complexities of several modulation/spreading schemes, we decided to use the pilot-assisted M-ary bi-orthogonal keying (MBOK) scheme [14]. The principle of MBOK can be explained with the help of Fig. 6, which shows the block diagram of the baseband transmitter. A data symbol, representing eight data bits, consists of three code sequences transmitted simultaneously: one data code sent along the Q-channel, and the other data code and the pilot code sent both in the I-channel. The eight possible data codes transmitted in I- and Q-channels are selected out of a set of 16-chip orthogonal Walsh codes. Two groups of 3-bit input data in the current symbol then determine which of the eight Walsh codes are sent in the I- and Q-channel, respectively. The remaining two bits of the eight-bit symbol determine the polarities with which the two data codes are transmitted. The predetermined pilot code sequence, again a 16-chip Walsh code, is always transmitted in the I-channel with positive polarity. The eight data codes plus the pilot code, since chosen from 16-bit Walsh codes, are thus mutually orthogonal. Table I shows the pilot code and the correspondence between the 3-bit input data and the eight 16-chip codes.

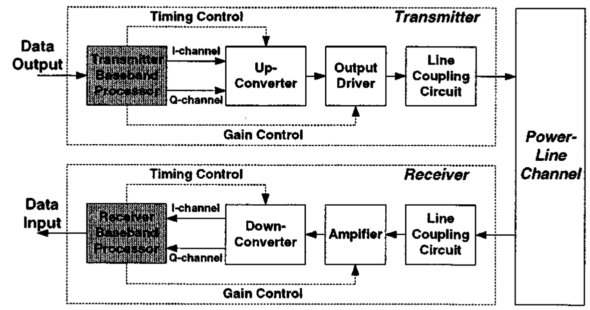

Fig. 5. Block diagram of a power-line communication system.

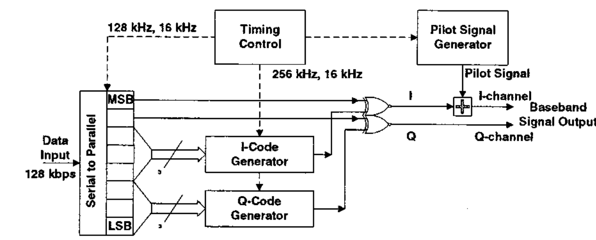

Fig. 6. Baseband transmitter architecture of the pilot-assisted MBOK modulation.

This scheme adopts a dedicated pilot code to convey channel information for use by the channel estimator in the receiver. Because the pilot code is transmitted with the data codes simultaneously through the channel, it suffers from the same distortion and phase change as the data codes. Thus, the receiver is able to come up with a very good estimate of the channel. Channel estimation can easily be accomplished through a complex correlator that extracts the phase of the received pilot signal.

Obviously, using a dedicated pilot code decreases power efficiency as well as communication efficiency of the system. The addition of the pilot code increases transmission power by 1.5 fold, which means a 1.8 dB performance degradation. Nevertheless, we note that this degradation is quite small compared to the dynamic range of channel attenuation, which can be as large as 45 dB [10]. Hence, we conclude that it is the channel condition rather than the transmission power level that determines

TABLE I
RELATIONS BETWEEN DATA BITS AND TRANSMITTED CODES

Data	Code
Pilot	+ − − + + − − + + − − + + − − +
000	+ + + + − − − − + + + + − − − −
001	+ − + − − + − + + − + − − + − +
010	+ + − − − − + + + + − − − − + +
011	+ − − + − + + − + − − + − + + −
100	+ + − − − − + + − − + + + + − −
101	+ − − + − + + − − + + − + − − +
110	+ + − − + + − − − − + + − − + +
111	+ − − + + − − + − + + − − + + −

communication quality. This justifies our use of pilot code for channel estimation.

B. Frame Structure

To limit transmission power within the frequency band below 450 kHz, the transceiver is designed to operate at a chip rate of 256 Kcps. This gives a symbol rate of 16 Ksymbol/s, and a raw channel data rate of 128 Kbps, since there are 16 chips in a symbol duration and a symbol represents eight input bits. However, in order to minimize the effect of sudden change of channel characteristics and lower the probability for a transmitted frame being hit by impulse noises, the frame length is made very short. In addition, short frames also make timing synchronization easier. Suppose the length of a data frame is short the timing error accumulated within it will be so small that it can be safely neglected. For example if a frame runs 1000 chips in length, assuming perfect synchronization at the beginning of a frame and the clock frequencies in transmitter and receiver have a mismatch of 100 ppm, there will be an accumulated timing error of about 0.1 chip at the end of this frame, which cannot seriously degrade system performance. Therefore, the timing synchronization problem is simplified to the issue of achieving accurate synchronization between the transmitter and the receiver at the start of each frame.

As shown in Fig. 7, a frame is composed only of 30 data symbols, preceded by a preamble section for frame detection and timing synchronization. The preamble is composed of eight periods of two spreading waveforms $C_a(t)$ and $C_b(t)$ transmitted in the I- and Q-channel, respectively. The waveform $C_a(t)$ is a 15-chip m-sequence PN code. The other spreading waveform $C_b(t)$ is derived by shifting $C_a(t)$ by a half period, i.e.,

$$C_b(t) = C_a(t - T/2)$$

where $T = 15T_c$ is the period of the spreading waveforms. The last period in the preamble is the Start-of-Frame Delimiter (SFD), having opposite polarity to previous seven cycles to mark the end of the preamble and the beginning of the data symbols. There is a short interval of one chip duration between the preamble section and the frame body, wherein no signal is transmitted. This guard interval prevents interference between the preamble and the frame body, and gives the receiver sufficient time to switch its state from preamble detection to data reception.

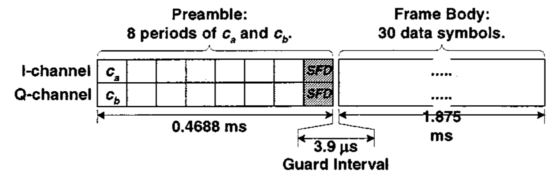

Fig. 7. Frame structure.

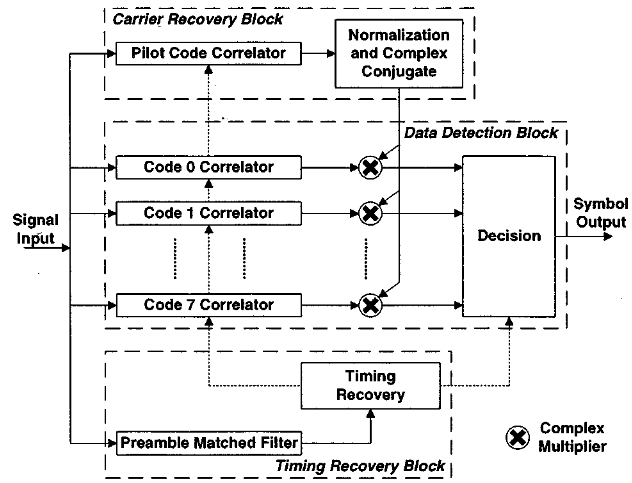

Fig. 8. Block diagram of the baseband receiver.

C. Receiver Design

At the receiver end, the received passband signal is sampled at 1.024 MHz, four samples per chip, and down-converted to baseband before entering the baseband processor. Fig. 8 shows the block diagram of the receiver baseband processor. There are three main functional blocks: the carrier recovery (channel estimation) block, the data detection block, and the timing recovery block.

To demonstrate the operation of demodulation in the receiver, suppose that a transmitter along a power line sends a symbol consisting of the pilot code and the ith data code in the I-channel, and the jth data code in the Q-channel. The corresponding baseband signal can be written as

$$C_p(t) + P_i \cdot C_i(t) + j \cdot P_j \cdot C_j(t)$$

where $C_p(t)$ represents the pilot code, and P_i, $C_j(t)$, P_j, $C_j(t)$ are the polarities and codes transmitted in the I- and the Q-channel, respectively. Then assuming perfect timing synchronization between the receiver and the transmitter and noiseless channel, the receiver picks up the signal

$$[C_p(t) + P_i \cdot C_i(t) + j \cdot P_j \cdot C_j(t)] \cdot e^{j\phi}$$

where ϕ is the total phase shift introduced by the transmit filter, power-line channel, and the receiver filter. This signal then enters eight data code correlators in the data detection block, as well as the pilot code correlator in the carrier recovery (channel estimation) block. After a symbol period, outputs of the ith and the jth data code correlators, and the pilot code correlator are $P_i \cdot e^{j\phi}$, $j \cdot P_j \cdot e^{j\phi}$, and $e^{j\phi}$, respectively. Because the data codes and the pilot code are mutually orthogonal, outputs from all other correlators are zero. The pilot code correlator output $e^{j\phi}$, which estimates the channel phase shift ϕ, will be taken complex conjugate. The result $e^{-j\phi}$ is a reference phase and can be used to com-

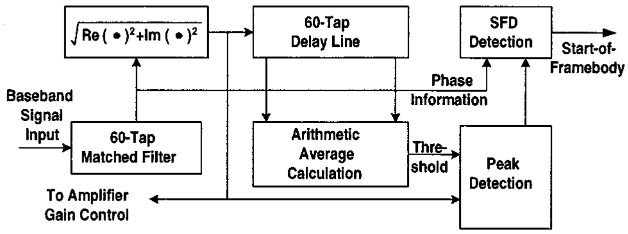

Fig. 9. Block diagram of the timing recovery circuit.

pensate for ϕ. Before entering the decision unit, outputs from the eight data code correlators are further multiplied by this reference phase. Thus, outputs of the ith and jth complex multipliers becomes P_i and $j \cdot P_j$, while others are still zero. The decision unit will recognize that, among all its inputs, the ith input has the largest magnitude in the real part and the jth input has the largest magnitude in the imaginary part. It can then be decided that the ith code is transmitted in the I-channel and the jth code in the Q-channel. Moreover, the polarities with which they are transmitted, i.e., $\mathrm{sgn}(P_i)$ and $\mathrm{sgn}(P_j)$, can also be determined.

D. Timing Synchronization and Frame Detection

Because the correlation properties of Walsh codes is different from those of pseudonoise spreading codes (e.g., m-sequences code), MBOK is quite sensitive to timing error between the transmitter and the receiver. Timing synchronization is thus crucial to the successful operation of this power-line communication system. Another important issue is the detection of a data frame among various noises in the power-line channel. A robust frame detection mechanism that can provide high detection probability and low false-alarm probability is required.

A detailed structure of the timing recovery block in the MBOK receiver is shown in Fig. 9. It has a matched filter matched to the waveforms $C_a(t)$ and $C_b(t)$ used in the preamble. This matched filter continuously reads in the baseband signal and calculates the correlation values between the incoming signal and the preamble m-sequence codes. If a preamble reaches the receiver, the magnitude of the complex correlation output from the matched filter will exhibit large periodical peaks; otherwise, only random noise signal can be observed at its output.

The output from the matched filter then feeds the peak detection circuit and a 60-tap delay line. In the peak detection circuit, the correlation output magnitude are checked if they are large peaks. First, a large peak must be a peak, which means that it has the largest correlation magnitude among neighboring seven correlation values, including itself. Secondly, a large peak must have a correlation magnitude larger than a threshold. This thresholding procedure reduces the number of false alarms by ruling out small peaks. The threshold value is determined by the arithmetic mean of the previous 60 correlation magnitude values, so the threshold value is adjusted dynamically according to channel condition. The arithmetic mean needed for thresholding is calculated in the arithmetic averaging circuit.

If large peaks periodically reappear, the phases of these peaks are checked for SFD in the SFD detector. A 180° phase change marks the SFD and it causes the timing recovery block to stop timing synchronization operation. Data detection block and carrier recovery block in the receiver are then triggered to process the following data symbols in this frame.

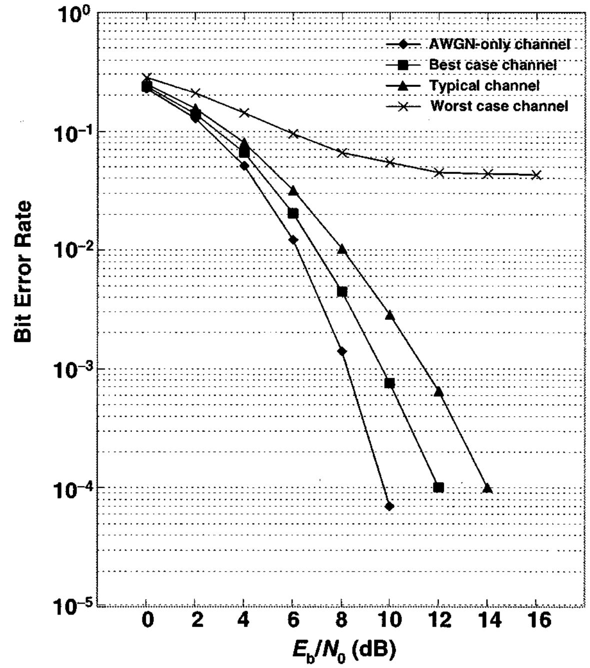

Fig. 10. Simulated performance of the baseband receiver with fixed-points data operations.

IV. Hardware Implementation and Measurements

A. Circuit Design

Most blocks in the power-line modem baseband receiver, including correlators, preamble matched filter, and timing synchronization circuit, work at the sampling rate of 1.024 MHz. On the other hand, complex phase rotation and data decision are executed at the symbol rate of 16 kHz. Hence, the requirement for speed is quite loose, while hardware complexity becomes the major concern. In the receiver, operations at symbol rate are done serially instead of in parallel, which greatly reduces the hardware complexity. Only one complex multiplier is needed instead of eight and in the decision unit two comparators, one for the real part and one for the imaginary part, will suffice.

Another important issue in circuit design is the word-length of the signals in the receiver. Minimizing word lengths of signals in the receiver will of course reduce the hardware complexity, but will induce more quantization noise and thus degrade receiver performance. A series of simulation was performed to find the proper word lengths of several crucial signals, such as receiver input, correlator output, matched filter input, and matched filter output. We found that for reliable performance under large Gaussian noise, 8-bit resolution is needed for the input signal and the correlator output. Moreover, the matched filter should adopt 5-bit input and 8-bit output. With word lengths of these important signals set, those for other signals can be accordingly found without further truncation in the following signal processing.

B. Performance Simulation

Fig. 10 shows the simulated BER performance of the proposed power-line modem under the aforementioned four channel scenarios: AWGN-only, best, typical, and worst. Results for BER less than 10^{-4} are not obtained due to simulation

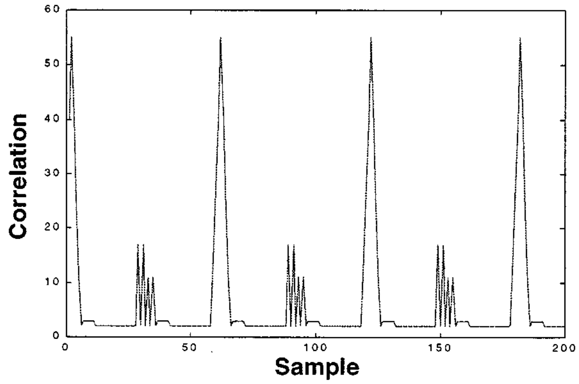

Fig. 11. Measured output waveform from the FPGA implementation of the matched filter.

time constraint. We note that the performance of this modem under a typical channel condition is 4 dB worse than that under an AWGN-only channel at a BER of 10^{-4}. This performance degradation is within our design margin and is acceptable. However, for the worst-case channel scenario the baseband receiver performs poorly and has a BER floor at about 4×10^{-2}. The reason for this is that under the worst-case channel, AWGN is not the dominant cause of errors while the continuous wave noise and impulse noise with large amplitudes overwhelm the finite dynamic ranges of the signals in the receiver, which are determined using the AWGN-only channel. Therefore, in order to circumvent this difficulty, signals in the receiver must be represented using more bits (wider word length). Since cost is one of our major concerns, we decided to tolerate this high data loss probability in the worst-case channel scenario and use the word lengths obtained previously in the final design.

C. Experimental Results

The power-line modem baseband transceiver was implemented on an FPGA chip. First the gate-level descriptions of the transceiver system were designed and then the detail circuit configuration was synthesized by a logic synthesis tool. The configuration file was down-loaded to the FPGA chip on a demo board. Testing and verification were done with the help of a logic analysis system. First, test patterns were generated on a PC with the proposed transmitter and power-line channel model program and then they were transferred to a digital pattern generator. The pattern generator fed the FPGA with desired signal while a logic analyzer monitored the outputs from the FPGA chip and recorded the waveforms.

During all verification procedures, the system operates at 5 MHz (five times the specified speed). The measured output waveform of the matched filter is shown in Fig. 11. Further measurements show that the remainder of the FPGA transceiver, including transmitter, matched filter, timing synchronization, and data decision, all function correctly.

V. Conclusions and Discussions

With proper communication techniques, one can use the power line as a communication medium for home automation and communication applications. To this end, an MBOK modulation scheme providing 128 Kbps data rate over the power line and its corresponding baseband processor is proposed. To facilitate the development of this processor, a power-line channel model is constructed for system performance evaluation and functional verification. The modulation scheme and hardware architecture are designed to combat the impairments of the hostile power-line channel and provide high-quality communication with minimum hardware. Real-time finite-precision FPGA implementation confirms that this pilot-assisted MBOK modem is feasible and cost-effective. We believe that the proposed modulation scheme and transceiver architecture provide a sound and solid foundation for reliable data communication through power lines.

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