A SPREAD SPECTRUM MODEM FOR USE ON ELECTRICAL POWER LINES

by

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ABSTRACT

The design, implementation, and testing of a spread spectrum modem for use on electrical power lines are described in this thesis.

It is shown that for data communication over power lines, spread spectrum signalling has performance advantages over signalling formats which use a narrow frequency band. Synchronization for the modem is based on the 60 Hz power line signal and significantly reduces the complexity of the modem. A spread spectrum transmitter and receiver are implemented and performance results on some representative power line circuits are presented.

The modem performed well and appears to be a viable candidate for data communications over power lines.
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1. INTRODUCTION TO POWER LINE CARRIER

1.1 Uses of Power Line Carrier (PLC)

Power line carrier involves transmission of data using lines whose primary purpose is the transmission or distribution of electric power.

1.1.1 Uses of High Voltage Lines

High voltage (HV) lines have been used to carry information as well as power for some time. In some cases the lines carry long distance telecommunication services such as voice or low bit rate data channels. Using the HV lines for communications is particularly attractive for developing countries which do not have widespread land lines or microwave links. In this case the desired information bandwidth is large and almost any number of channels made available could be put to use. However, the channel characteristics (see Section 2.1.4) normally limit the information bandwidth.

Another service which is sometimes implemented using HV lines as a communications channel is power system telemetry. In this case data on generating plants, remote substations, and the HV lines themselves are sent over the HV lines to a central monitoring station. This central station is then able to monitor the state of the power system.

1.1.2 Uses on Distribution Lines

The low voltage distribution wiring has also been used for communications. PLC on distribution wiring is used to provide meter reading,
load shedding, and other services [1]. The motivation for automated meter reading is to reduce the costs associated with manual reading [2]. In addition, remote meter reading allows price rationing, thereby controlling electricity demand during peak periods [3]. This controlling of electric power reduces the requirements for auxiliary generators and hence costs. Similarly, automatic shedding of non-essential loads reduces auxiliary generator requirements and potential brown-out problems.

Caldwell has done a cost/benefit analysis of these services for the London area [4]. His figures indicate that automatic load shedding and meter reading are economically viable provided that meter reading capabilities are used to provide price rationing (time of day tariffs).

Both of these services require the communication of data between the customer's premise and the local Hydro office. Meter reading requires electric power consumption data to be transmitted from the customer to the Hydro office, while load shedding requires shedding commands to be transmitted in the opposite direction.

The data rates required for these services can be estimated as follows. For meter reading (time of day billing), assume four billing periods per day. A "smart" meter keeps a running total of the number of kilowatt-hours used during each time-of-day period. Once a month a single message is sent to the Hydro office containing the previous month's electricity usage. Figure 1.1 shows a possible format for this message which contains 124 bits and includes a start field, an address field for over one million customers, an error control field, the electricity usage, and a stop field.
124 BITS/MESSAGE

Figure 1.1 A monthly meter reading message.
Assuming one substation for every 100,000 customers the data rate required for meter reading is calculated using equation 1.1.

\[
R_m = 10^5 \text{ customers} \times \frac{124 \text{ bits}}{\text{customer-month}} \times \frac{1 \text{ month}}{2.6 \times 10^6 \text{ sec}} = 4 \text{ bit/sec} \quad (1.1)
\]

For loading shedding assume each shedding command affects ten customers (except special emergency shedding commands which would affect many hundreds of customers). Also assume that one command (msg) is needed to turn non-essential loads off, another message to turn them back on, and that this on-off cycle occurs an average of twice a day. Finally, let the message be the same length as the message shown in Figure 1.1. then, the data rate required for load shedding is given by (1.2).

\[
R_s = 10^5 \text{ customers} \times \frac{4 \text{ msg}}{10 \text{ customers-day}} \times \frac{124 \text{ bits}}{\text{msg}} \times \frac{1 \text{ day}}{10^5 \text{ sec}} = 40 \text{ bits/sec} \quad (1.2)
\]

These data rates are estimates only but are close to the data rates used on some commercial PLC communication systems [1].

1.1.3 Uses on Intrabuilding Lines

Table 1.1 lists some of the communication services which could be provided within a home, office or factory [5]. There are a number of communication channels which could be used to provide these services such as CATV cable [6], telephone lines, infrared radiation [7] or power line carrier (PLC).

The advantages of using PLC are that the channel is already in place, it has a standard mechanical interface in the wall socket and power plug, and
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Table 1.1 Intrabuilding Communications Services
it runs almost everywhere inside and outside homes, offices, and factories. Its disadvantages are that it has limited bandwidth compared with fibre optics or CATV cable, it is noisy, and its impedance, attenuation, and topology are not well defined. (See Section 2.1).

1.2 PLC Modem Requirements

Modem operational parameters of interest include data rate and error rates. Another factor which impacts modem design is the switching scheme used. These matters are discussed in this section.

1.2.1 Data Rates Required for PLC

The requirements of a PLC system will vary depending on the type of services carried. Data rates of a few bits/sec would satisfy the requirements of meter reading while digital voice would require over 10 Kbits/sec. Table 1.2 lists the data rates of some PLC services which could be provided. Notice that many of the services require fewer than 60 bits/sec, a few require rates near 1Kbit/sec, while few others require data rates in excess of 100 Kbit/sec.

Table 1.2 lists some services only. Many others exist and even more would be proposed in the future. It seems then that any PLC system must be flexible enough to handle a wide variety of information rates.

Some of the data rates listed in Table 1.2 need clarification. Digital voice can be encoded in a variety of ways; companded PCM requires 64 Kbits/sec [15] while adaptive delta-modulation can provide adequate quality using 16 Kbits/sec [16]. Some of the phoneme-based speech synthesizers
<table>
<thead>
<tr>
<th>Service</th>
<th>Data Rate</th>
</tr>
</thead>
<tbody>
<tr>
<td>Meter reading</td>
<td>~4 bits/sec (See Section 1.1.2)</td>
</tr>
<tr>
<td>Pwr systems telemetry</td>
<td>~10 bits/sec (See Section 1.1.1)</td>
</tr>
<tr>
<td>Load shedding</td>
<td>~40 bits/sec (See Section 1.1.3)</td>
</tr>
<tr>
<td>Voice (digital)</td>
<td>~10 Kbits/sec</td>
</tr>
<tr>
<td>Voice (analog)</td>
<td>~4 KHz</td>
</tr>
<tr>
<td>Bldg security</td>
<td>~10 bits/sec</td>
</tr>
<tr>
<td>Bldg intercom</td>
<td>~4 KHz or 10 Kbits/sec (per channel)</td>
</tr>
<tr>
<td>Stereo distribution</td>
<td>~40 KHz or 1 Mbit/sec</td>
</tr>
<tr>
<td>Terminal-to-computer link</td>
<td>~50 bits/sec</td>
</tr>
<tr>
<td>Computer-to-terminal link</td>
<td>~100 bits/sec</td>
</tr>
<tr>
<td>Computer-to-computer link</td>
<td>~300 bits/sec to 64 Kbits/sec</td>
</tr>
<tr>
<td>Video distribution</td>
<td>~6 MHz or 80 Mbits/sec</td>
</tr>
<tr>
<td>Appliance control</td>
<td>~10 bits/sec</td>
</tr>
<tr>
<td>Environmental management</td>
<td>~10 bits/sec</td>
</tr>
</tbody>
</table>

Table 1.2 Data rates required to implement some PLC services.
require approximately 1 Kbit/sec [17], [18]. Hence, the 10 Kbits/sec quoted for voice and intercom is a compromise. Analog voice as currently provided on both telephone and radio voice channels uses approximately 4 KHz of bandwidth.

Services such as building security, appliance control, and environmental management can be implemented in a variety of ways. Typically these services might be implemented as follows. A maximum of $1024=2^{10}$ devices might be specified which could be used as sensors for security and environmental monitors or controlled switches for appliances. The message format would contain 10 bits of address, 8 bits for command or status information, an error control field of eight bits, and 4 bits for start, stop, and overhead. A typical message would therefore contain around 30 bits. Some of the devices such as temperature sensors or appliance switches would not be accessed more than a few times a day. Other devices such as a security sensor may be polled every few seconds or may independently send alarm messages. Depending on the number of active devices we could expect a few messages per hour in a lightly loaded single family home to over ten messages per second in a factory with many sensors. The figure of 10 bits/sec given in Table 1.2 is a compromise between these extremes.

Stereo distribution requires an analog bandwidth of 20 kHz per channel. Digital stereo with high fidelity must be sampled at 40 kHz, with 12 bits/sample, for a total of about 800 Kbits/sec on two channels.

Data communications between computers and terminals is highly variable. A skilled operator can type approximately 10 characters per second (120
words/min). Using 8 bits/character gives a peak data rate from terminal to computer of 80 bits/sec. However, this rate is rarely sustained, except possibly in word processing. Thus, 50 bits/sec seems to be a safe estimate.

Computer to terminal information rates are also highly variable. Individuals can read at about 250 words/min (≈ 150 bits/sec) but rarely sustain that rate during computer transactions. The rate of 100 bits/sec represents about 180 words/min. Computer to computer communication can be at almost any rate. 300 bits/sec used to be standard over telephone lines but recently 1200 bits/sec or higher has become commonplace. Some specialized data communication services such as packet switched networks offer data rates of 64 Kbits/sec.

Finally video distribution requires 6 MHz of bandwidth using the NTSC standard while digital video can range from 1 Mbit/sec for picturephone [19] to 80 Mbits/sec for 8 bit PCM video [20].

1.2.2 Error Rates Tolerable for PLC

The error rates tolerable for the services listed in table 1.1 are difficult to estimate. Customers using meter reading and load shedding equipment seem satisfied with a bit error rate of $10^{-4}$ [21]. Modem manufacturers try to deliver a bit error rate of better than $10^{-6}$ while users of digital voice tolerate an error rate of $10^{-2}$ [22].

A PLC modem might provide the user with the capability to trade off error rates against data rates [23]. One way to accomplish this objective is to make the code rate variable, which involves a scheme to control undetected
errors by trading off the number of information bits against the number of error control bits [23].

The type of errors (bursty, random, etc) must be known if the most effective error control strategy is to be selected [23]. Unfortunately, there is not much data available on PLC error statistics.

1.2.3 Switching Alternatives for PLC

The switching requirements for a PLC communications system are varied. Some services, such as computer to terminal communications, are effectively handled by packet switching. Other services which include file transfers are efficiently handled using a circuit switched network.

Sometimes point-to-point communications is required, such as when a message from an individual customer's hydro meter is sent to the billing office. In other cases broadcast capability is required, for example when a load shedding command is to be sent to several hundred customers. Finally, some analog services, such as home intercom, may use the PLC channel. It appears that a multipurpose PLC modem should allow for different switching alternatives.

1.3 Outline of the Thesis

Chapter two is a discussion of previous work which impacts spread spectrum modem design. The communications characteristics of the power system are examined in Section 2.1. In Section 2.2 PLC signalling techniques which have already been tried by others are summarized. Section 2.3 is an introduction to spread spectrum communications.
Chapter three deals with the system design of the spread spectrum power line modem. Sixty Hertz synchronization and its implications are outlined in Section 3.1. Section 3.2 explains the transmitter and receiver structures required to implement a direct sequence spread spectrum PSK modem. The choice of the bandwidth selected for communication is explained in Section 3.3. The network which coupled the modem into the power system is discussed in Section 3.4 while Section 3.5 discusses the code generator.

Chapter four documents the modem implementation. Included are block diagrams, circuit schematics, and circuit descriptions.

Chapter five contains the results of tests carried out with the modem. In Section 5.1 background information on modem testing is given. Section 5.2 describes the results of tests done in a controlled environment. Section 5.3 and Section 5.4 are the results of tests done on some representative power circuits. A summary and discussion of the results is included in Section 5.5.

Conclusions and suggestions for future work are given in Chapter six.
2. REVIEW OF PREVIOUS WORK

2.1 Power System Transmission Characteristics

The transmission characteristics of power transmission and distribution systems have been, and continue to be, the subject of extensive study. Items of interest include the propagation of power system transients, the propagation of conducted power line noise, and the propagation of power line carrier signals within the power system. The factors which influence the transmission characteristics of the power system are discussed in this section.

2.1.1 Power systems topology

Electrical power is distributed from generating plants to users over a series of distinct networks. These networks operate at different voltages and are interconnected by transformers.

First the high voltage network carries power from generating plants to electrical substations. Figure 2.1 is a map, published by B.C. Hydro, of the high voltage lines existing or planned in B.C. as of 1974. These lines vary greatly in length from a few to hundreds of kilometres. Figure 2.2 is the electrical equivalent of one high voltage line showing its generator, branches, loads (which vary with time and frequency), and the line itself.

Next, the distribution network delivers electric power from a substation to a number of distribution transformers. Figure 2.3 is a map of part of this distribution network for Vancouver. These distribution lines vary in length, too. In addition they have a large number of branches and loads, presented by the distribution transformers, with complex impedances. These loads, like the lines, have impedances which vary with frequency. The load
Figure 2.1 The high voltage network in British Columbia.
Figure 2.2 The electrical equivalent of a high voltage line [24].
Figure 2.3 The distribution network.
impedances, however, also change with time as electrical loads (such as home appliances) are switched on and off. In some cases individual distribution lines connect to each other forming a loop back to the distribution transformer. (These loops provide redundancy. If a single fault occurs at any point in the loop users will still receive power.) Figure 2.4 is the electrical equivalent of part of the distribution network.

Residential wiring is similar to the distribution network [25]. Each distribution transformer supplies power for a number of houses. Figure 2.5 shows a typical connection from a distribution transformer to the houses. Although a single-phase transformer is shown, three-phase transformers are also used. In turn, each house contains many individual loads.

Figure 2.6 shows the wiring in a typical house. Notice that each household load may connect to either or both phases. Also, there are few restrictions on the actual number of loads which may be connected at a given time in a given household. (There are some restrictions - each branch must contain no more than twelve different loads. In addition, some circuits, including those which supply appliances, are further restricted as to loading [26].)

Office and industrial wiring resemble residential wiring in that a variety of loads can be connected to an individual circuit although three-phase power and different voltages are sometimes used.

A wide variety of loads can be connected to the power line. On the high voltage and distribution lines these loads are usually restricted to transformers, which can present virtually any impedance, and power factor
Figure 2.4 The electrical equivalent of a distribution line.
Figure 2.5  Residential wiring between houses and the distribution transformer.
Figure 2.6 Typical residential wiring.
correction loads. Residential and commercial wiring, however, can supply loads which are time-varying in number and type. These loads may be inductive, capacitive, resistive, or nonlinear. They may draw currents ranging from milliamps to hundreds of amps. Finally, except for regulations covering new equipment [27], they can conduct and/or radiate any amount of electrical noise back on to the power line.

2.1.2 Power line noise

Power lines carry electrical noise along with the 60 Hz power. This noise is generated from a large number of sources in a large number of ways.

Power line noise can be short lived. This transient noise can arise from events such as lightning strikes (Figure 2.7) or powering on an inductive load (Figure 2.8).

Power line noise can be transient in the short term and periodic in the long term. This type of noise is typically generated phase controlled loads (see Figure 2.9). Power line noise can be coupled in externally from radio and T.V. stations. Power line noise can be narrowband (Figure 2.10). This type of noise is often produced by digital equipment where many internal devices will turn on and off (producing transient noise) in response to some master clock [28]. Finally, power line noise can be wideband. This type of noise can arise from spark gaps or corona discharges [29].

Although it is instructive to look at sources and characteristics of power line noise, it is the noise seen by a receiver which affects the performance of PLC systems. The transmitted and received noise are not the
Figure 2.7 Transient recorded during lightning storm.

Figure 2.8 Transient recorded during starting of a furnace blower.
Figure 2.9 Transient power line noise.

Horizontal = 2 ms/div

Vertical = 5 volts/div
Horizontal = 10 KHz/cm
Vertical = 10 dB/cm

Figure 2.10 Narrowband power line noise.
same. The difference arises from the transmission characteristics of the power system.

Smith [30] has done a survey of power line noise coupled into a 50Ω receiver. His results are shown in Figure 2.11 for various environments. These results show average levels only, although short term variations of 10 dB were observed. Notice how the noise levels vary with the environment. The rural levels are almost 30 dB below the urban office levels! Notice also that the noise is not white but falls off at about 20 dB/dec. The roll-off results from both lower generated noise and higher attenuation at high frequencies (see Section 2.1.4).

Assuming most loads generate noise, we would expect that power line noise would decrease during periods of inactivity (typically at night). Similarly, periods of heavy electrical use (around dinner time) should also have increased noise. In other words, background noise should correlate well with the electrical load profile.

Some power line noise measurements were recorded, over a 24-hour period, in Room 325 of the Hector MacLeod building, U.B.C. Also, power line noise, again over a 24-hour period, was measured in Suite 206 – 3680 West 7th Avenue, Vancouver, B.C. (A residential complex). The background noise in the Hector MacLeod building was fairly consistent varying less than 2 dB over the measurement period. However, the background noise in the residential complex varied by more than 6 dB during a 24-hour period. The highest noise levels occurred at night around 11:00 p.m. while the lowest levels were present in the early morning.
Figure 2.11  Power line conducted noise.
In addition to background noise levels, electrical interference from common household appliances were also measured. The measurements indicate that most loads do not significantly add to the background noise. The worst noise offenders were the vacuum cleaner and the light dimmer and these added less than 3 dB to the background noise. Finally, very little appliance noise couples across from one phase to the other.

2.1.3 Signal propagation at low frequencies

Power lines contain many impedance discontinuities which cause signals propagating down the line to reflect and backscatter [31]. The backscattered signals will combine with the forward signal to create areas of increased and decreased signal strength. In other words standing waves will be set up along the power line.

These transmission line effects become pronounced when the impedance discontinuities are separated by more than approximately an eighth of the signal's wavelength. Therefore, the frequency at which transmission line effects become noticeable will depend on the environment through which the signal passes. For example, if a propagation velocity of free space is assumed then a signal's wavelength is given by

\[ \lambda = \frac{c}{f} \]  

(2.1)

where \( \lambda \) is the signal's wavelength, \( C \) is the speed of light, and \( f \) is the signal's frequency.

In residential wiring individual circuits rarely run longer than three hundred metres, so from (2.1), PLC signals below 100 KHz will not suffer from transmission line effects. On the distribution network circuits can run tens
of kilometres so signals above 10 kHz will be affected. Finally, on the high
electricity power transmission network, power lines can run for hundreds of kilo-
metres. Transmission line effects on lines this long occur even for the 60
Hz power signal.

Ignoring transmission line effects permits a simple model for PLC sig-
nal propagation at low frequencies. At low frequencies power line conductors
have a very small series impedance. For example common household wiring
(number twelve gauge) has an equivalent series resistance of .006 Ω/m and a
series inductance of approximately .43 µH/m depending on the material in
close proximity to the conductors [32]. At a frequency of 100 kHz, and a
length of 30 metres, the series impedance of the conductors is given by:

\[ |Z_{CONDUCTOR}| = \sqrt{R_s^2 + X_s^2} = \sqrt{(.006 \times 30)^2 + (.43 \times 10^{-6} \times 30 \times 2 \pi \times 10^5)^2} = 1 \Omega \]

This impedance is less than one-tenth of the minimum load impedance which can
be connected across a residential circuit. Thus the conductor impedances
\((Z_C)\) are small compared to the load impedances \((Z_L)\) connected between them.
This means that at low frequencies the equivalent load impedance presented to
a PLC transmitter or receiver, is approximately equal to the parallel com-
bination of all the loads on the line (see Figure 2.12).

The actual power delivered to the receiver from the transmitter will
depend on this equivalent line impedance \((Z_L)\). (At low frequencies it does
not matter where along the line the transmitter and receiver are placed. The
power delivered to the receiver is independent of position.) The equivalent
line impedance will be time varying, as loads are turned on and off, and
Figure 2.12(a) Power line impedance at low frequencies.

Figure 2.12(b) Equivalent power line impedance at low frequencies.
frequency varying, depending on the type of loads connected at any given instant.

Some measurements of the line impedance were made at different times in the Hector MacLeod building, U.B.C. The results of these measurements are shown in Figure 2.13. Notice that the line impedance varies with both time and frequency. At frequencies above 100 kHz there is increased variation in $Z_L$ because transmission line effects become noticeable. At one point $Z_L$ dropped off sharply at 400 kHz. This decreased impedance was due to reflections which occurred at that instant of time and frequency. (Reflections, or standing waves, will cause the line impedance to vary with position along the line [33].)

The attenuation of the PLC signal between the transmitter and receiver depends on the equivalent line impedance. Because this impedance varies with time and frequency the attenuation will also vary. Figure 2.14 shows the attenuation between a transmitter and receiver measured at different times within a building.

2.1.4 Propagation at high frequencies

With increased frequency the skin effect increases the per unit resistance and inductance of the line causing its characteristic impedance and its attenuation to increase [34]. The characteristic impedance of a transmission line depends on many factors, such as conductor size, spacing, height above ground, dielectric, etc. For power lines the characteristic impedance can vary from 400 Ω in widely separated overhead lines to 20 Ω for underground lines [35].
Figure 2.13 The measured power line impedance.
Figure 2.14  Typical transmission versus frequency characteristic [38].
When transmission line effects are included, the transmission characteristics of the power system become very difficult to predict because of branches in the line and the time and frequency varying loads. Work has been done on modelling the high voltage and distribution networks at high frequencies [35], [36]. However, even though some general RF characteristics of the power system are known there are still many uncertainties [37].

The impedance seen by power line transmitters and receivers will vary with time and position as well as frequency. In general, however, the transmission line will transform "distant" load impedances to the characteristic line impedance and loads in close proximity will therefore have the biggest impact [34].

Attenuation along the power line is generally quite low because the conductors are large. On overhead lines the attenuation can be less than .1 dB/Km [35]. However, wideband signal attenuation due to capacitive loads [38] and distribution transformers [39] will be present (see Figure 2.15).

Finally, narrowband signal fades can occur because of standing waves and multimodal propagation [35]. These narrowband fades will be position and time dependent (see Figure 2.16).

2.1.5 Received signal to noise variations

The performance of any communication system is dependent on the signal to noise ratio (SNR) available at the receiver. In power line communication systems the channel characteristics discussed in this section make the received SNR highly variable and unpredictable.
Figure 2.15 Channel characteristics of a channel loaded with some specific devices.

Figure 2.16 Narrowband channel drop-outs due to standing waves.
The SNR will vary with time as loads are connected and disconnected. More loads generally result in more attenuation and more noise. Thus better performance would be expected at night when fewer loads are connected across the line.

The SNR will vary with frequency. The frequency dependence arises both from loads and the line itself. The line, because of the skin effect, attenuates high frequencies more than low frequencies. Different loads will affect PLC signals differently. Resistive loads, such as heating elements, will attenuate signals evenly. Inductive loads, such as motors, will attenuate low frequencies more than higher ones. Capacitive loads, such as power factor correction capacitors, will attenuate high frequencies more than lower ones.

The SNR will vary with position; this position dependence is due to a number of factors. First, narrowband fades due to reflections and multimodal cancellation are position dependent. Second, signal losses increase as the receiver is moved farther from the transmitter. Third, background noise will be worse close to a noisy load.

In summary, the signal level at a receiver depends on the following factors: 1) how much transmitter power is coupled into the line (the degree of coupling depends on the transmitter's source impedance and the instantaneous line impedance presented to the transmitter); 2) how much power is coupled into the receiver (this coupling again depends on the receiver's input impedance and the line impedance at the receiver); and 3) how much signal power is attenuated between the transmitter and receiver which is influenced by the line and its loads. The noise level, and power spectrum,
at the receiver will depend mostly on the type of loads connected nearby. Generally, the noise is approximately +10 dBm, into a 5 ohm load at 10 kHz, and falls off at approximately 20 dB per decade.

2.2 PLC Signalling Techniques

A large number of signalling techniques for power line carrier have been tested [40]. Information has been sent at baseband, at very low frequencies, at audio frequencies, and at frequencies above 100 kHz. The variety of existing systems indicates that there is not yet a consensus on the best system for power line carrier signalling. This section discusses the most common power line carrier systems together with their advantages and disadvantages.

2.2.1 Very low frequency signalling

Very low frequency (VLF) signalling is a data communications system where signalling is done at a very low rate, usually a fraction of the power-line frequency. VLF signalling circumvents many of the channel anomalies mentioned in Section 2.1 by ensuring that no high frequency signals are generated. Thus, transmission line and topological effects can be ignored and the power distribution grid appears as a broadcast channel with negligible transmission line losses. With VLF signalling, only varying channel noise and load impedances need be considered.

Forrest and Gray propose a VLF signalling scheme in which every fourth zero crossing of the distribution voltage waveform is either distorted (i.e. transmitting a one) or undistorted (transmitting a zero) [41]. This distort-
tion is accomplished by purposely connecting a load at the appropriate zero crossings, or by purposely generating a switching transient (see Figure 2.17). Receivers connected to the line monitor the line zero crossings for these switching transients (see Figure 2.18) and assemble messages.

VLF signalling, however, does have drawbacks. First, only very low data rates can be transmitted (less than 100 bits/sec). Generally, for meter reading, this data rate restriction is not unmanageable since meter data rates are on the order of a few hundred bits per month per household. Thus, tens of thousands of households could be accommodated.

A second difficulty with VLF signalling is the uplink channel, or the channel from the customer to the hydro office. The technique used by Forrest only provides a simplex link from the office to the customer. Thus, VLF signalling is applicable only to load shedding and other signalling techniques must be employed to enable two-way communications.

2.2.2 Ripple control

Ripple control has been used for power line carrier since 1929 [42]. Since that time ripple control has evolved into a number of related systems. Ripple control is basically ASK signalling with a carrier frequency between 110 Hz and 750 Hz. These frequencies are low enough so that transmission line effects can be safely ignored.

Figure 2.19 shows a ripple control transmitter. The data to be transmitted controls the firing of the SCR's in the static frequency converter. The audio frequency used is chosen to lie between two of the 60 Hz harmonics to minimize noise. The series resonant circuit \((C_g \text{ and } L_g)\) blocks the 60 Hz
Figure 2.17 A VLF transmitter and the resulting distribution line voltage.

Figure 2.18 A VLF receiver.
Figure 2.19 A ripple control transmitter.

Figure 2.20 A ripple control receiver.
while the parallel resonant circuit \((C_p, L_p)\) couples the audio power to the line. The transformer provides isolation. The parallel resonant circuit also filters the higher audio harmonics generated by the frequency converter.

The ripple control receiver is shown in Figure 2.20. A coupling network similar to the transmitter passes the carrier, rejects the 60 Hz and isolates the receiver from the power line. A non-coherent AM detector follows and passes the received data to the decoding and control circuits.

The advantages of ripple control are its simplicity and its low carrier frequency. The low carrier frequency means that transmission line effects can be ignored. Ripple control also has disadvantages, including a low data rate (about 2 bits/sec), simplex only transmission, and very high transmitter requirements (greater than 50 KVA!). The high transmitter power is required because the power line noise is very high at low frequencies.

2.2.3 Baseband signalling

Baseband signalling is used to transmit and receive data in the vicinity of the zero crossings of the 60 Hz power [43]. During this interval both the power voltage and noise are low and the line (at least between transformers) looks like a d.c. link. Noise is low in this interval because motors and switches draw (and hence generate) little power. However, some loads such as switching power supplies and data processing equipment (which operate asynchronously from the 60 Hz) will generate noise in this interval.

Figure 2.21 shows a typical baseband transmitter and receiver. In this system coupling to the line is easy. A d.c. connection is made to the line.
Figure 2.21 A baseband transmitter and receiver.
during the zero-crossing interval. Beside simple coupling, the system has other advantages. Two-way communication is possible and the $T_x$ power required is low (because the noise is low). Finally, high data rates can be achieved (more than 10 Kbits/sec).

The biggest disadvantage of this system is that the signals propagate poorly through the distribution transformer. For this reason baseband systems normally operate only within a building.

2.2.4 Two-way signalling near 10 KHz

This class of signalling techniques attempt to overcome the disadvantage of ripple control (see Section 2.2.2). First, the power line noise is smaller at 10 KHz than 400 Hz. Second, the power line has a higher impedance. These two factors combine to reduce the transmitter power required. A lower transmitter power also makes multiple transmitters feasible with the result that two-way communication is possible. The actual signalling can be ASK [44], PSK [45], or FSK [43], all of which have been tried.

One disadvantage of these systems is that they still require moderate transmitter power, typically 400 watts, which makes the transmitter expensive. Another disadvantage is that the signals are still attenuated somewhat by the distribution transformers and power factor correction capacitors.

2.2.5 Two-way signalling near 50 KHz

At high carrier frequencies the transmitter power required is around 1 watt [44] and coupling to the power lines can be done simply [9]. These facts make the communication equipment cheaper than most of the lower fre-
quency alternatives. However, transmission line effects (see Section 2.1.4) become more pronounced so performance is less predictable.

Even at these frequencies there is wide disagreement on which frequencies are best suited to power line carrier. Frequencies from 10 KHz to over 100 KHz have been tried [43], [44], and [46].

2.3 Spread Spectrum Communications

Spread spectrum communications is a signalling technique which increases the bandwidth (spreads the spectrum) of a signal to make it less susceptible to interference [47] and narrowband fades [48]. The spreading process can be accomplished by one of four techniques or combinations of them.

Chirp spread spectrum systems [49] sweep a carrier, usually linearly, within a frequency band \((f_L, f_H)\). A 'one' could be transmitted by sweeping from \(f_L\) to \(f_H\) while a 'zero' could be transmitted by sweeping from \(f_H\) to \(f_L\).

Frequency hopped spread spectrum systems [50] shift the carrier frequency to a number of discrete frequencies within a frequency band \((f_L, f_H)\). The shift from one frequency to another is usually done in a pseudo-random fashion.

Time hopped spread spectrum systems [50] turn the carrier on and off in a pseudorandom order. Time and frequency hopped systems are often combined into a hybrid system [51].
2.3.1 Direct sequence spread spectrum communications

Figure 2.22 shows a direct sequence spread spectrum (DS-SS) communications system. In DS-SS the information bit-stream d(t) is first multiplied by a high rate pseudorandom bit stream c(t) before being PSK modulated by the carrier k(t). Similarly, the receiver, after conventional demodulation, remultiplies by c(t) to recover the information bit stream.

To gain more insight into the spreading and despreading processes Figures 2.23 and 2.24 show the various signals in the time and frequency domains. In this example the information rate = 1/T₁ bits/sec and the chip rate (i.e., the pseudorandom bit stream rate) = 1/T_c = 10x1/T₁. In addition, the carrier frequency is f_c = 1/T_c. Also included is wideband interference.

In this case the processing gain = G_p = 10 log₁₀ (T₁/T_c) = 10 dB. Notice that the SNR is -5 dB at RF but +5 dB at baseband because of the processing gain.

2.3.2 Performance characteristics of DS-SS

Spread spectrum signals are insensitive to narrowband fades which occur in PLC as a result of reflections and multimode propagation (see section 2.1.4). Spread spectrum communication is not degraded severely by narrowband fades because the RF energy is spread over a large number of frequencies. Thus a narrowband fade will remove only a small portion of the RF signal so that the total energy is not significantly reduced [48].

Spread spectrum signals are insensitive to narrowband interference which can be present in PLC systems (see section 2.1.2) [52]. Spread spectrum signals can be successfully demodulated even with a negative received signal-to-noise ratio. Thus, with a power density below that of the noise
Figure 2.22 A direct sequence spread spectrum communications system.
Figure 2.23 DS-SS PSK time domain waveforms.
Figure 2.24  DS-SS PSK frequency domain power spectrums.
floor SS systems can co-exist with other narrowband systems [53], such as analog voice intercom.

Finally, more than one spread spectrum signal may be transmitted simultaneously without excessive mutual interference. (The interference increases as the number of signals increase.) Other SS signals appear as wideband noise to an SS receiver whose despreading code does not match those of other signals; codes are chosen to have small cross-correlations [54]. Thus some security of the information sent over power lines ensues, since information cannot be demodulated unless the code is known.

2.3.3 Disadvantages of spread spectrum communications

The wide spectral occupancy of spread spectrum signals causes some difficulties. The power line has both amplitude and phase characteristics which are frequency dependent. In narrowband communications these non-linearities appear constant within the band of interest. However, wideband signals will suffer distortion unless these effects are equalized [55].

Spread spectrum signals must be demodulated synchronously. The codes in the receiver and transmitter must match or the processing gain of the link will be reduced [56], with the result that code synchronization must be acquired and kept for the duration of the message.
3. PLC MODEM SYSTEM DESIGN

3.1 Block diagram of a Direct Sequence Spread Spectrum PLC Modem

The block diagrams of the transmitter and receiver are shown in figure 3.1.

In the transmitter the data is first spread by the multiplier and then up-converted by the modulator. The resulting radio frequency (RF) signal is then amplified and coupled onto the power line. (The sequence of spreading and up-converting can be reversed without affecting the modem characteristics).

The code generator which is discussed in Section 3.5, generates the high rate pseudorandom sequence which multiplies (spreads) the data stream. The spreading factor (processing gain) is about 400 times (26 dB) and is discussed in Section 3.2.

The PSK modulator converts the baseband signal to RF. The RF frequency band chosen, which is discussed in Section 3.3, is a compromise between attenuation at high frequencies and noise at low frequencies. The transmit amplifier must be wideband to pass the spread RF signal and have linear phase to minimize distortion. Finally, the line coupling network, discussed in Section 3.4, must provide isolation, surge protection, impedance matching and 60 Hz blocking.

The receiver is similar in structure to the transmitter, with additional system units for synchronization at the carrier, code, and bit level, and for detection. The details of synchronization are discussed in Section 3.2.
Figure 3.1 Transmitter and receiver block diagrams for the DS-SS -PLC modem.
3.2 Synchronization

3.2.1 Code synchronization in spread spectrum communications

Code synchronization is critical to spread spectrum communications. If synchronization is lost there is no processing gain and hence no interference rejection. Figure 3.2 shows the autocorrelation function of a pseudorandom (m-sequence) which is N bits long. In the example shown N = 15. The autocorrelation function is at a maximum at zero time shift and then falls off sharply as the time shift increases. Note that if the time shift is greater than one bit period then all correlation is lost. In general, synchronization between the transmitted code sequence and the local replica, generated at the receiver, should be within one half of a bit.

A number of techniques have been developed to acquire and maintain this synchronization. One procedure is to generate a local replica of the code and then "slide" this replica and the received signal by each other until they crosscorrelate. Because of the large number of positions to try, the sliding correlator is slow to acquire synchronization. As a result modifications are often added to reduce synchronization delay. Sometimes a short code is transmitted first to acquire synchronization and then the long code is used to transmit the data. (The code must be as long as possible to spread the data uniformly). Other times special codes with peculiar cross-correlation properties which speed synchronization are used [50].

Other techniques such as matched filter receivers or sequence estimators are quicker but have other disadvantages, including sensitivity to noise and high implementation costs.
Figure 3.2 Autocorrelation of an m-sequence of N bits (N=15).
3.2.2 Code synchronization by zero crossings

Whichever process is used, synchronization slows down and complicates the transmission of data through a spread spectrum modem. For this reason synchronization should be simplified if possible. Some applications such as satellite communications make use of universal timing [50]. In this case all transceivers operate off the same timing signal and only propagation delays determine the phase of the received signal.

A universal timing signal is already present in power line carrier communications - the a.c. power frequency. Code synchronization can be achieved automatically if all transceivers synchronize to the 60 Hz zero crossings as shown in Figure 3.3. In this example the code is fifteen bits long and starts on every positive-going zero crossing of the power line voltage. Receivers generate local replicas of the code which, ignoring propagation delays, matches the received code generated by the same zero crossings at the transmitter.

3.2.3 Carrier and bit synchronization

Synchronization by zero crossings can be extended to both carrier and bit synchronization as well as code synchronization. Synchronizing the carrier, code, and bits gives a coherent communications system which performs better than a non-coherent system [57]. Because zero crossings provide this coherence the cost and complexity disadvantages normally associated with coherent communications may be substantially reduced.

Bit synchronization can be accomplished by limiting the data rate to an integer multiple of the power line frequency. This may seem restrictive at
Figure 3.3 Code synchronization by zero crossings.
first but most of the services listed in Chapter One require data rates of
less than 60 bits/sec. These services can therefore be accommodated by one
bit every power line cycle. Higher data rate services can also be accommo­
dated because the standard data transmission rates of 300, 1200, 2400 and
9600 bits/sec are all multiples of 60 Hz.

Similarly carrier synchronization can be acheived by making the carrier
frequency a multiple of the code rate and hence a multiple of the power line
frequency. Again this may seem restrictive but there are many multiples to
choose from. Thus the RF spectrum can be placed wherever desired. (See
Section 3.3) Figure 3.4 illustrates the bit, code, and carrier synchroniza­
tion. The bit frequency is 60 bits/sec, the code length is 15 bits, and the
carrier frequency equals the code rate. Note that all the timing (synchroni­
zation) is referenced to the positive going zero crossings of the power line
voltage. The bottom waveform in Figure 3.4 is the spread spectrum signal
which results from multiplying the data, the code, and the carrier together.

3.2.4 Implications of power line synchronization

Power line synchronization is simple to implement and hence is an
inexpensive way to provide synchronization to all transceivers. However,
power line synchronization does have a number of disadvantages.

Propagation delays of both the 60 Hz and spread spectrum signals will
cause phase uncertainties at the receiver. Propagating at the speed of light
the SS signal and the zero crossings of the 60 Hz line voltage move about 300
m every microsecond. Inside a building this delay is negligible because
Figure 3.4 A coherent, line synchronized, spread spectrum system.
transmission distances are short. However, on the distribution or high voltage networks the distance between transceivers may be many kilometres.

As an example, consider the case of a receiver located at a substation receiving a message from a transmitter located at a house 10 Km away. The receiver sees a zero crossing and starts its local code generator. The same zero crossing then propagates 10 Km to the house where the transmitter senses it and starts its local code generator. The 10 Km delay is about 30 microseconds, assuming signals propagate at close to free space velocity (the speed of light). The transmitter's signal then propagates the 10 Km back to the receiver making a total delay of about 60 microseconds. If the code rate is 10 Kbits/sec then each chip (bit of the code) has a 100 μsec duration. Thus a delay of 60 μsec is six tenths of a chip. From Figure 3.2, 6/10 of a chip reduces the processing gain by more than fifty percent. If the house had been 20 Km away there would be no processing gain at all.

Fortunately the transceivers are not mobile so it is possible to pre-compensate for the propagation delay. Precompensation would be done at the transmitter as follows: For a message travelling down the line in the direction of power flow, no compensation is necessary because both the power and the SS signal will travel at close to the same velocity. For a message travelling up the line opposite the direction of power flow the transmitter will start the code and the data bits slightly before the zero crossing. The distance between the transmitter and the receiver will determine the amount of precompensation such that the start of the code will reach the receiver at the same time the zero-crossing does.
Another source of phase uncertainty exists because wall sockets, and therefore the transceivers attached to them, can be connected to different phases of the power system. In normal residential wiring, each house is connected to one of the three phases of the distribution network. In addition, each wall socket in the house connects to one of the two phases. Thus there are six possible phases as shown in Figure 3.5. In addition to the six possible phases, each transceiver can be plugged into one of two polarities because the plugs are not normally polarized.

The resulting phase ambiguity can be resolved by making the code repeat six times within each cycle. In this way the code will restart at each of the six possible phases. It doesn't matter in this instance to which phase the transceivers are connected. The polarity ambiguity can be resolved by differentially encoding the data [57]. In differential encoding, the received polarity is unimportant. Only the difference in polarity between the current bit and the previous bit determines the data. For example, a "one" would be encoded by making the current bit opposite from the previous one while a "zero" would be the same as the previous bit. A small penalty is paid when the method is used because each received error usually generates two decoded errors.

A final source of phase uncertainty is phase noise (jitter). The phase noise will cause the zero crossings to jitter, with the result that codes in the transmitter and receiver will move with respect to each other. The magnitude of this jitter will determine the effective processing gain of the system. For example, if the jitter is less than one tenth of the chip duration, the effective processing gain will not be significantly less than the
Figure 3.5  The six possible power line voltages.
maximum gain obtainable with no jitter. If, however, the jitter is about equal to the chip duration then the effective gain will be about one half of its maximum value.

Jitter in the zero crossing comes from three sources. Amplitude noise (AM) on the 60 Hz line voltage will convert to phase noise (PM) when passed through a threshold detector as shown in Figure 3.6. This AM to PM conversion can be significantly reduced by filtering the line voltage to remove the amplitude noise before passing it to the threshold detector.

Phase noise is also produced at generating stations [32]. The composite load presented to the generators will vary as individual loads are turned on and off. These varying loads cause the generators to vary their speed of rotation. As the speed of rotation changes so does the instantaneous frequency as shown in Figure 3.7. This instantaneous frequency variation is equivalent to phase noise. This source of phase noise will cause jitter but all transmitters and receivers will encounter the same jitter. Thus, the zero crossings at different transceivers will not move with respect to each other and as a result the processing gain will not be reduced.

The final source of jitter is difficult to eliminate. This jitter is also caused by varying loads but will be different for different transmitters and receivers. The mechanism shown in Figure 3.8 represents two equal power line segments \((Z_L)\) terminated by two different loads \(Z_{A_1}\) and \(Z_{A_2}\). It should be noted that the phases of the 60 Hz voltages \(\angle V_1\) and \(\angle V_2\) depend on the loads \(Z_{A_1}\) and \(Z_{A_2}\). These loads will be different and time varying so the
Figure 3.6 Zero crossing jitter due to noise on the 60 Hz line voltage.
Figure 3.7  Instantaneous frequency variations in electrical generators.
Figure 3.8  The effect of unequal loads on zero crossing jitter.

\[ \angle V_1 = \angle A_1 - \angle \theta_L \]

\[ \angle V_2 = \angle A_2 - \angle \theta_L \]
60 Hz phases at the receiver and transmitter - and hence the zero crossings - will vary with respect to each other. Fortunately, the line impedances ($Z_l$) are much smaller than the load impedances, so the phase changes are very small.

Figure 3.9 is an oscilloscope photograph which illustrates the zero crossing jitter. Notice that peak one-sided jitter is about 10 µsec. The amount by which the jitter reduces the processing gain depends on the code rate. If the code rate is 10 Kbits/sec (chip duration = 100 µsec), the jitter will be less than 10% of the chip duration, and the effective gain will not be significantly reduced. If, however, the code rate is 100 Kbits/sec (chip duration = 10 µsec) the jitter will be close to 100%, and the effective gain is reduced by about one half.

The theoretical processing gain is the ratio of spread to unspread bandwidth, or the ratio of code rate to data rate as shown in equation (3.1).

$$\text{Processing gain} = \frac{R_{\text{code}}}{R_{\text{data}}}$$

(3.1)

For a fixed data rate, the processing gain increases as the code rate increases. However, as the code rate increases the jitter becomes a larger percentage of chip duration, and the effective processing gain decreases. At some point, as shown in Figure 3.10, a maximum processing gain is reached where any further increase in code rate causes a net decrease in processing gain.

It is possible to solve for this maximum processing gain. Assume $N$ code chips per 60 Hz cycle ($R_{\text{code}} = N \times 60$) and a measured peak jitter of $J$
Horizontal = 20 µsec/div.

Figure 3.9  Measured zero crossing jitter.
Figure 3.10  Maximum processing gain.
sec. Redrawing Figure 3.2 in the vicinity of zero time shift (see Figure 3.11) gives an equation for the effective processing gain as a function of $N$ and $J$:

$$G_{\text{eff}}(J,N) = G_p[1 - J \times N \times 60] \quad (3.2)$$

The maximum processing gain ($G_p$) is directly proportional to the chip rate and hence directly proportional to the number of chips per 60 Hz cycle ($N$). Thus, $G_p$ can be written as follows:

$$G_p = K_p \cdot N \quad (3.3)$$

Combining (3.3) and (3.2) gives:

$$G_{\text{eff}}(T,N) = K_p \cdot N - K_p \cdot 60 \cdot J \cdot N^2 \quad (3.4)$$

Differentiating (3.4) with respect to $N$ gives:

$$\frac{\partial G_{\text{eff}}(T,N)}{\partial N} = K_p - K_p \cdot 60 \cdot J \cdot 2 \cdot N \quad (3.5)$$

Equating (3.5) to zero and solving for $N$ yields:

$$N_{\text{opt}} = \frac{K_p}{K_p \cdot 60 \cdot 2 \cdot J} = \frac{1}{120 \cdot J} \quad (3.6)$$

Equation (3.6) shows that if the jitter is small one can select a large value for $N$, in which case a high code rate and therefore a large processing gain results. Inserting (3.6) back into (3.4) yields:

$$G_{\text{eff}}(T)_{\text{max}} = \frac{K_p}{120J} - \frac{K_p \cdot 60 \cdot J}{120 \cdot 120 \cdot J^2} = \frac{K_p}{120J} (1 - \frac{1}{2}) \quad (3.7)$$

Thus at the maximum gain, the effective gain is down by 1/2 (3 dB) from its theoretical value as a result of the jitter.
Figure 3.11 Effective processing gain.
A measured value for $J$ is about 10 $\mu$sec. Using this value in (3.6) gives $N_{\text{opt}} = 830$. At the minimum data rate of one bit every 60 Hz cycle, the theoretical maximum processing gain from (3.1) is:

$$G_{\text{p,max}} = \frac{R_{\text{code}}}{R_{\text{data}}} = \frac{N_{\text{opt}} \times 60}{1 \times 60} = N_{\text{opt}} = 830$$  (3.8)

From (3.7) the effective maximum processing gain is one half of this, or $G_{\text{peff,max}} = 400$. Converting this gain to a power ratio gives:

$$G_{\text{eff,max}} (\text{dB}) = 10 \times \log_{10} (G_{\text{eff,max}}) = 26 \text{ dB}$$  (3.9)

Therefore, the maximum processing gain is limited to about 26 dB when zero crossings of the power line voltage are used for synchronization. In this application 26 dB is adequate, and line synchronization would therefore appear to be viable.

3.3 Frequency Band of Operation

The transmission characteristics of the power system include high noise power at low frequencies and high attenuation at high frequencies (see Section 2.1). These characteristics indicate that the frequency band between 10 kHz and 100 kHz is most suitable for PLC. There are, however, special cases.

On the HV network the conductors are large and therefore have lower attenuation at high frequencies than distribution or intrabuilding wiring. Also, most PLC communications on the HV network do not pass through power
transformers, which are a source of high frequency attenuation. Therefore, PLC on the HV network is generally at a higher frequency.

Within a building, distances are so short that attenuation - even at higher frequencies - is not too large. It is also feasible to install high frequency bypass capacitors between the phases so transformer losses will not occur at high frequencies. These factors allow PLC communications at higher frequencies within a building.

In the case of narrowband PLC, the choice of carrier frequency is critical. It would not be advisable to set the carrier frequency to 10 kHz, for example, because of interference from OMEGA, a marine location service which uses high power transmitters operating at 10 KHz [58]. It would be better to set the carrier between two of the 60 Hz harmonics to reduce the noise level [42].

For SS communications a wide frequency band is used and the above considerations are not as critical. For simple synchronization it is important instead to make the carrier frequency a multiple of the line frequency (see Section 3.2). To make things even simpler the carrier frequency should be equal to the code rate. In that way only one frequency need be synthesized.

Using these constraints the modem will have the characteristic listed in Table 3.1. The processing gain listed in Table 3.1 is an approximation which takes into account the gain reduction due to jitter (see Section 3.2). Using these frequencies the modem will occupy the spectrum as shown in Figure 3.12.
<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data rate ($D_T$)</td>
<td>60 bits/sec.</td>
</tr>
<tr>
<td>Code rate ($C_T$)</td>
<td>14 Kbits/sec.</td>
</tr>
<tr>
<td>Carrier frequency</td>
<td>14 KHz</td>
</tr>
<tr>
<td>Processing gain</td>
<td>$10 \log_{10} \left( \frac{C_T}{D_T} \right)$ - Jitter reduction = 20 dB.</td>
</tr>
</tbody>
</table>

Table 3.1 Modem characteristics.
Figure 3.12  Spectral occupancy of the spread spectrum signal.
3.4 The Line Coupling Network

The line coupling network connects the modem to the power system. It provides isolation, surge protection, 60 Hz blocking, impedance matching and low insertion loss of the PLC signal.

Isolation must be provided because most power plugs are not polarized, that is the plug can be inserted in two ways and the ground wire will be connected to either of the two modem wires as shown in Figure 3.13. Isolation is most simply achieved magnetically with a transformer.

Surge protection is necessary because high voltage transients are present in the power system and would otherwise damage the modem.

The coupling network must simultaneously block the 60 Hz power and pass the PLC signal. This can be accomplished by a high pass filter. The filter's cut-off frequency should be as low as possible so it will pass most of the PLC signal while still providing sufficient attenuation of the power signal.

Finally, the coupling network must match impedances between the modem and the power line, to maximize the signal to noise ratio. Unfortunately, this matching is difficult to achieve because both the noise and power line impedances are highly variable. (See Section 2.1) Figure 3.14 shows the situation at the transmitter.

The voltage on the line \( V_L \) will contain both the transmitted signal \( V_S \) and noise \( V_N \). Figure 3.14 can be further simplified by combining the two noise sources and their impedances as shown in Figure 3.15. In this case the signal power \( P_S \) is given as
Figure 3.13  Polarity ambiguity in modem connection.
Figure 3.14   Impedances seen by a PLC transmitter.
Figure 3.15  Equivalent impedances seen by a PLC transmitter.
\[ P_s = \frac{V_s^2}{Z} = \left( \frac{V_T \cdot \frac{Z_N}{Z_T + Z_N}}{Z_T \cdot \frac{Z_N}{Z_T + Z_N}} \right)^2 \]  

\hspace{1cm} (3.10)

While the noise power \( P_N \) is given by

\[ P_N = \frac{V_N^2}{Z} = \left( \frac{V_N' \cdot \frac{Z_T}{Z_T + Z_N}}{Z_T \cdot \frac{Z_N}{Z_T + Z_N}} \right) \]  

\hspace{1cm} (3.11)

Thus the SNR is

\[ \text{SNR} = \frac{P_s}{P_N} = \left( \frac{V_T}{V_N'} \right)^2 \cdot \left( \frac{Z_N}{Z_T} \right)^2 \]  

\hspace{1cm} (3.12)

Differentiating (3.12) with respect to the transmitter's impedance \( Z_T \) and equating this to zero gives the \( Z_T \) which maximizes the transmitted SNR.

\[ \frac{\partial \text{SNR}}{\partial Z_T} = -\frac{1}{2} \times \left( \frac{V_T}{V_N'} \cdot \frac{Z_N}{Z_T} \right)^2 \times Z_T^{-3} = 0 \]  

\hspace{1cm} (3.13)

Unfortunately, (3.13) is only zero if \( Z_T \) is zero. This value can not be obtained practically but equation (3.13) indicates that \( Z_T \) should be made as small as possible to make the SNR as large as possible.

A similar analysis can be done at the PLC receiver and leads to the same conclusions. The receiver's impedance should be as small as possible. Figure 3.16 shows the final configuration of the coupling network.
Figure 3.16  The coupling network.
3.5 The Code Generator

The code generator provides a bit stream $k(t)$ which must have a number of characteristics. First, the code should be part of a family so that a transmitter can send to a specific receiver by generating the receiver's code sequence. Each receiver has a unique code sequence which is a member of the family. Codes of this type allow privacy. An unauthorized receiver cannot decode the message unless it knows the particular code used. (However, a user with suitable processing equipment could learn the code identity).

Second, the cross-correlations between the code sequences should be low. Codes of this type allow code division multiple access (CDMA) in that a number of messages can be transmitted simultaneously. Any given receiver can only decode (despread) a message whose code matches. All other messages — provided their codes have low cross-correlation with the receiver's code — will appear as background noise.

Finally, the codes should have a low autocorrelation except at zero time shift. (See Figure 3.17.) This really means the codes should be as random as possible. Truly random code sequences will have no discrete lines in their power spectrum, making them the least susceptible to fades and interference. In other words, random codes have no concentrations of signal power which could be removed by a fade or interfered with by a jamming signal. Unfortunately, a truly random signal could not be reproduced at the receiver so pseudo-random codes must be used as a compromise.

M-sequences have this ideal autocorrelation property. Unfortunately, m-sequences have poor cross-correlation properties because members are just time-shifted versions of each other. Thus, all cross-correlation functions
Figure 3.17 Ideal autocorrelation function of the spreading code.

Figure 3.18 Cross-correlation between two m-sequences.
between m-sequences will have a "peak" where the time shifts cancel. (See Figure 3.18.) This peak in the cross-correlation function can cause difficulties when a receiver is attempting to acquire synchronization. The receiver may falsely "lock" to the incorrect transmitter, in a SDMA system, because the sliding correlator will lock onto the first correlation peak it encounters. Because the PLC modems use universal timing they can not falsely acquire synchronization. Therefore, the PLC modems can use simple m-sequences.
4. PLC MODEM IMPLEMENTATION

4.1 The Transmitter

The PLC transmitter accepts data which is then spread, modulated, amplified, and coupled into the power system. The circuits which implement these functions are discussed in this section.

4.1.1 The transmitter block diagram

Figure 4.1 is a block diagram of the PLC spread spectrum transmitter. Synchronization for the modem is based on the power line frequency and allows a simple implementation for the code generator which is discussed in Section 4.1.4. The data spreader and binary phase-shift-keyed (BPSK) modulator are also examined in Section 4.1.4. The transmit amplifier, built from an integrated circuit (IC) developed for the audio industry, is examined in Section 4.1.5. The power supply and protection circuits are discussed in Section 4.1.2 while the coupling network is detailed in Section 4.1.6.

4.1.2 The power supply

The circuit diagram for the power supply is shown in Figure 4.2. The transformer provides isolation and two phases of low voltage alternating current (AC). The bridge rectifier and capacitors $C_1$ and $C_2$ generate positive and negative unregulated direct current (DC) voltages. These two voltages are regulated by $U_1$ and $U_2$ which are fixed five volt positive and negative series-pass regulators. Capacitors $C_3$, $C_4$, $C_5$, and $C_6$ provide additional filtering.
Figure 4.1 The transmitter block diagram.
Figure 4.2 The power supply.

F1 = 1/2 Amp
C1, C2 = 2200 μF @ 16V
C3, C4 = 10 μF @ 25V
D2 = GE Transorb #V130LA20A
U1 = μA7805
U2 = μA7905
T1 = Hammond #161K12 (6.3V to centre tap)
D1 = EBR #VS048 (2 Amps)
Also shown in Figure 4.2 is the protection circuit. It consists of a fuse and a transorb which protect the modem from the high voltage transients which occur on the power line.

4.1.3 The code generator

Figure 4.3 is a detailed block diagram of the code generator. The code generator accepts the power line voltage as its input and generates the pseudorandom sequence and the carrier, which are both synchronized to the power line frequency.

The zero-crossing detector generates a pulse (ZERO) on every positive going zero-crossing of the power line voltage. This pulse is then used to synchronize both the carrier and the code sequence. Figure 4.4 is the schematic of the zero-crossing detector [59], while Figure 4.5 is an oscilloscope photograph showing its performance.

The clock generator is driven by a colour-burst (3.58 MHz) crystal and synchronized by the ZERO pulse. Figure 4.6 is the schematic of the clock generator and Figure 4.7 is its timing diagram. The circuit delivers a carrier which is synchronized to the power line voltage.

The 3.58 MHz oscillator [60] generates a high frequency clock which is then divided down to produce the carrier. The two dividers together divide by 256 so the carrier frequency is about 14 KHz. Line synchronization is provided by the CLR signal, which is derived from the ZERO pulse as shown in the timing diagram (Figure 4.7). The CLR pulse clears the two counters so they always start counting from zero at every zero-crossing of the power line voltage. Thus, the carrier goes through about 233 cycles before being
Figure 4.3 The code generator block diagram.
Figure 4.4 The zero crossing detector.
Horizontal = 2 ms/div.
Top trace = power line voltage.
Bottom trace = output of zero crossing detector.

Figure 4.5 Output of the zero crossing detector.
Figure 4.6  The clock generator schematic.
Figure 4.7 The clock generator timing diagram.
restarted by the CLR pulse at the next zero-crossing.

The code generator accepts the carrier signal and the ZERO pulse. It then generates the code sequence which is used to spread the data. The code generator is simply an m-sequence generator [61] which is restarted by the ZERO pulse so that code synchronization is simplified (see Section 3.2.2).

Figure 4.8 is the schematic of the code generator and Figure 4.9 is an oscilloscope photograph showing its operation. The two flips flops (74LS74) format the ZERO pulse so it can be used by the eight bit shift register (74LS299). Taps one and seven of the shift register are brought out to an exclusive NOR gate (74LS266). The gate's output then feeds back to the input of the shift register. This feedback arrangement generates a pseudorandom m-sequence which is 127 (2^7-1) bits long. Because there are 233 carrier cycles within every 60 Hz cycle the code sequence repeats almost twice between each restart pulse.

4.1.4 The data spreader and BPSK modulator

The data spreader is simply a digital multiplier which multiplies the low speed data by the high speed code. This multiplying process spreads the spectrum of the data. The multiplier can be implemented by an exclusive NOR gate as shown in Figure 4.10.

Also shown in Figure 4.10 is the BPSK modulator. It too can be implemented by an exclusive NOR gate. The carrier in this case is a square wave so the BPSK modulator is sometimes called a biphase modulator.
Figure 4.8 The code generator.
Horizontal = 2 ms/div.
Top trace = restart pulse
Bottom trace = the spreading code
Code rate = 14 Kbits/sec.
Code length = 127 bits

Figure 4.9 Operation of the code generator.
Figure 4.10 The data spreader and BPSK modulator.
Figure 4.11 is a spectrum analyzer photograph which shows the spectral occupancy of the data (a) after it has been spread by the multiplier and (b) after it has been modulated by the BPSK modulator.

4.1.5 The transmit amplifier

The transmit amplifier accepts the modulated data \( M_{DATA} \) and delivers an amplified differential version to the coupling network. The actual power delivered to the coupling network depends on the line impedance, which is time varying. Under normal circumstances, however, the transmitted power is about .5 watts. Figure 4.12 is a schematic of the transmit amplifier.

4.1.6 The line coupling network

The line coupling network is shown in Figure 4.13. The circuit consists of an audio matching transformer, which provides isolation and impedance matching. Also included is a third order Butterworth high pass filter, which blocks the 60 Hz power signal while passing the high frequency PLC signal.

The transmit amplifier has a low output impedance (See Section 4.1.6) which is lowered still further by the matching transformer. The equivalent source impedance, on the secondary side of the transformer, is less than one ohm. This low source impedance is required for a high transmitted SNR (See Section 3.4).

The high pass filter is designed for a nominal cut-off frequency of 6 KHz. This frequency is adequate to pass most of the PLC spectrum (see Figure 3.12) while providing more than 100 dB of attenuation at 60 Hz. However, the
Figure 4.11(a) Frequency spectrum of the spread data ($s_{DATA}$).

Figure 4.11(b) Frequency spectrum of the modulated data ($M_{DATA}$).
Figure 4.12 The transmit amplifier.
Figure 4.13  The line coupling network.

C5,C6=13.2μF  (250V)
L1 = 2.5 mH   (2A)
T2=100Ω:3.3Ω  (1.5W)
line impedance will vary (see Figure 2.13) so the cut-off frequency will vary as well.

Figure 4.14 and Table 4.1 illustrate the transfer function of the coupling network. The load impedance is varied with frequency to simulate the power line impedance which varies with frequency. The primary side voltage ($V_p$) is kept constant to simulate the signal which would come from the low impedance transmit amplifier. Notice that the signal level is flat between about 6 KHz and over 100 KHz. The low frequency losses are due to the high pass filter while the high frequency attenuation is due to transformer loss.

<table>
<thead>
<tr>
<th>FREQ (KHz)</th>
<th>$V_p$ (volts p-p)</th>
<th>$V$(volts p-p)</th>
<th>$R_L$(Ω)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>10</td>
<td>.5</td>
<td>1</td>
</tr>
<tr>
<td>6</td>
<td>10</td>
<td>1.6</td>
<td>2</td>
</tr>
<tr>
<td>10</td>
<td>10</td>
<td>1.8</td>
<td>3</td>
</tr>
<tr>
<td>20</td>
<td>10</td>
<td>1.6</td>
<td>7</td>
</tr>
<tr>
<td>30</td>
<td>10</td>
<td>1.6</td>
<td>10</td>
</tr>
<tr>
<td>40</td>
<td>10</td>
<td>1.6</td>
<td>15</td>
</tr>
<tr>
<td>60</td>
<td>10</td>
<td>1.6</td>
<td>20</td>
</tr>
<tr>
<td>100</td>
<td>10</td>
<td>1.6</td>
<td>40</td>
</tr>
<tr>
<td>300</td>
<td>10</td>
<td>1.5</td>
<td>90</td>
</tr>
<tr>
<td>600</td>
<td>10</td>
<td>1.3</td>
<td>100</td>
</tr>
<tr>
<td>1,000</td>
<td>10</td>
<td>1.0</td>
<td>100</td>
</tr>
</tbody>
</table>

Table 4.1 Transfer function of the coupling network.

Figure 4.15 is an oscilloscope photograph of the PLC signal at the input (top) and the output (bottom) of the coupling network. The load impedance was set here to ten ohms as an approximation of the line impedance in the frequency band of the PLC signal.
Figure 4.14 Transfer function of the coupling network.
Horizontal = 100 $\mu$sec/div.

Top trace = $A_{\text{DATA}}$

Bottom trace = $L_{\text{DATA}}$

Line impedance = 10\,\Omega

Figure 4.15  Data transmission through the coupling network.
The output signal is only slightly distorted. The distortion is a result of the high pass filter in the coupling network which removes frequency components below 5 KHz. In addition there is negligible phase shift through the network. Figure 4.16 is a spectrum analyzer photograph which shows the spectral occupancy of the signal delivered by the transmitter.

4.2 The Receiver

The PLC receiver accepts the PLC signal from the power line. The signal is then amplified, demodulated, despread and finally detected. The circuits which implement these functions are discussed in this section.

4.2.1 The receiver block diagram

Figure 4.17 is a block diagram of the PLC spread spectrum receiver. The circuits which implement the protection network, coupling network, power supply, zero crossing detector, clock generator and code generator are similar to those found in the transmitter (see Section 4.1) Only the operating voltages are different: ±15 volts for the receiver and ±5 volts for the transmitter. The higher voltages in the receiver are required by the data desprredder (see Section 4.2.3). Figure 4.18 is a schematic of these circuits.

The receive amplifier is descussed in Section 4.2.2. This circuit amplifies the received PLC signal before passing it on to the demodulator and data desprredder, which are examined in Section 4.2.3. Finally, Section 4.2.4 concerns the data detector which generates the received data ($R_{DATA}$).
Vertical = 10 dB/cm.
Horizontal = 10 KHz/cm.
Line impedance = 10Ω

Figure 4.16  The spectrum of the transmitted signal.
Figure 4.17 The receiver block diagram.
Figure 4.18  Circuits similar to the PLC transmitter.
4.2.2 The receive amplifier

The receive amplifier is a differential amplifier which boosts the received signal \( R_{PLC} \) level before passing it on to the multiplier. The schematic diagram of this amplifier is given in Figure 4.19.

4.2.3 The data demodulator and despreader

Once the received signal has been amplified it must be demodulated and despread before being passed to the data detector. The amplified signal \( R_A \) is demodulated by multiplying it by the locally generated carrier.

\[
R_{DEM} = R_A \times R_{CAR} \tag{4.1}
\]

\( R_{CAR} \) is the locally generated carrier and \( R_{DEM} \) is the demodulated signal.

Next the demodulated signal is despread by multiplying it by the locally generated code sequence \( R_{CODE} \).

\[
R_{DES} = R_{DEM} \times R_{CODE} \tag{4.2}
\]

\( R_{DES} \) represents the despread signal. These two multiplications are shown in Figure 4.20.

Because the amplified signal \( R_A \) is analog, both the multipliers in Figure 4.20 must be four-quadrant analog multipliers. These are more costly and less accurate than an exclusive OR-gate (XOR) which can multiply two digital signals.

However, combining (4.1) and (4.2) results in a triple multiplication in which two of the terms \( R_{CAR} \) and \( R_{CODE} \) are digital signals. By changing the order to multiplication, which is commutative, one of the analog
Figure 4.19  The receiver amplifier.

\[ R_A = \frac{R_2}{R_1} = \frac{R_{PLC}}{R_{PLC}} \]
Figure 4.20 The data demodulator and despreader.
multipliers can be replaced by a digital one as shown in Figure 4.21. Figure 4.22 is the schematic of this simplified data demodulator and despreader.

The performance of the data demodulator and despreader is shown in Figure 4.23 and Figure 4.24. Figure 4.23 is an oscilloscope photograph showing the inputs and output of the analog multiplier. The top trace is the transmitted data \(T_{\text{DATA}}\). The next trace is the output of the XOR gate, which is a product of the locally generated carrier and code. The third trace is the amplified signal \(R_A\), while the bottom trace is the output of the multiplier - the demodulated and despread signal \(R_{\text{DES}}\). Figure 4.24 is an expanded view of Figure 4.23 at the point where the transmitted data changes polarity. This shows more clearly the operation of the analog multiplier.

4.2.4 **The data detector**

The data detector is of the integrate-and-dump variety. The block diagram of the detector is shown in Figure 4.25 while its schematic is detailed in Figure 4.26.

The integrator integrates the despread signal \(R_{\text{DES}}\) over one bit period. This signal is then applied to the threshold detector (comparator) the output of which is sampled at the end of the bit period. Figure 4.27 shows the performance of the data detector. The top trace is the despread signal \(R_{\text{DES}}\). The next trace is the ZERO signal, or the sampling point. The next trace is the output of the integrator. The bottom trace shows the output of the sampler, which is the received data signal \(R_{\text{DATA}}\).
Figure 4.21 A simplified data demodulator and despreader.
Figure 4.22 Schematic of the simplified data demodulator and despreader.
Figure 4.23  Operation of the data demodulator/despreader.

Top trace = $T_{DATA}$
Upper middle trace = CODE×CARRIER
Lower middle trace = $R_A$
Low trace = $R_{DES}$

Horizontal = 2 ms/div.
Figure 4.24 Expanded view of the data demodulator/despreader.

Top trace = $T_{DATA}$
Upper middle trace = CODE x CARRIER
Lower middle trace = $R_A$
Low trace = $R_{DES}$
Horizontal = 200 µsec/div.
Figure 4.25 The data detector.
Figure 4.26 The data detector schematic.
Top trace = $R_{DES}$
Upper middle trace = ZERO
Lower middle trace = $R_{INT}$
Low trace = $R_{DATA}$

Figure 4.27 Operation of the data detector.
5. MODEM PERFORMANCE

5.1 Test Procedure

Because the transmission characteristics of the power system vary over time, position, and frequency it is difficult to determine how well a PLC modem will work in any given situation. One approach to the testing and comparison of PLC modems would be to construct a communications channel with the transmission characteristic of a typical power line. All modems could then be tested on this channel and compared against each other. Unfortunately, a suitable channel model is not available. An alternative test procedure, which would serve well to define the general performance characteristics of the PLC modem, would involve actual operation on representative power line circuits.

In this regard, the spread spectrum PLC modem was exercised in three separate surroundings. First, the modem was tested on a communications channel which had a constant impedance, constant attenuation, and additive white Gaussian noise. Although this channel is not typical of actual power line transmission characteristics, this test does facilitate a comparison of the actual modem performance against the theoretically predicted performance. Any significant differences between the two would indicate that the modem was operating improperly.

Second, the modem was operated in a large multi-use building where a wide variety of loads were present, including industrial machinery, computers and computer terminals, and office equipment. This environment approximated that of industrial and office buildings.
In the third set of experiments, the modem was tested in an apartment complex of approximately thirty individual apartments. An apartment building this size provides a typical residential environment — with more electrical loads than a single family dwelling but fewer than a large high rise apartment complex.

The results of these tests are presented in Sections 5.2, 5.3, and 5.4. Section 5.5 summarizes and discusses the results.

5.2 Performance in a Controlled Environment

Figure 5.1 shows the equipment used for tests with additive white noise. To test the synchronization procedure both the transmitter and receiver are synchronized by the power line zero crossings. However, the PLC signal is sent over a separate channel. The channel in this case consists of a load resistor, which roughly approximates the power system impedance, and an additive white noise source.

In addition to the channel, the transmitter, and the receiver, other equipment is necessary for the test. A data source is needed to generate pseudorandom data bits \( T_{\text{data}} \) at the rate of 60 bits/sec. (In other words, one bit every power line cycle.) The data error counter compares the received data bits \( R_{\text{data}} \) against the transmitted bits and increments an error counter whenever there is a difference.

Because data errors occur randomly, the error rate is difficult to measure precisely. For example, a string of \( 10^4 \) data bits may have ten errors within the first one hundred bits sent. If the measurement is terminated after one hundred bits the error rate would be \( 10^{-1} \). If, however, the
Figure 5.1  The white noise test.
measurement is made over $10^4$ bits then the error rate would be $10^{-3}$. For this reason it is important to allow the modem tests to run for a long period of time so that the error rate can be measured accurately. In the case of the PLC modem all tests were run until close to 100 errors were accumulated to ensure an accurate $P(\xi)$ reading. The true RMS voltmeter is used to measure the power of the PLC signal and the noise which are present at the receiver.

The theoretically predicted performance of an integrate-and-dump detector operating in an additive white noise environment is

$$P(\xi) = \frac{1}{2} \times \text{erfc} \left( \frac{E_b}{\sqrt{N_o}} \right)$$  \hspace{1cm} (5.1)

where $P(\xi)$ is the probability of a received bit being detected incorrectly, $\text{erfc}(\cdot)$ is the error function complement, $E_b$ is the energy in each data bit as measured at the receiver, and $N_o$ is the noise power spectral density as measured at the receiver [62].

The probability of error is measured by the data error counter and is the ratio of the number of bits received in error relative to the total number of bits sent. The received bit energy is given by

$$E_b = (v_{R_{\text{RMS}}}^D)^2 \times T_b = (v_{R_{\text{RMS}}}^D)^2/f_b$$  \hspace{1cm} (5.2)

where $v_{R_{\text{RMS}}}^D$ is the RMS voltage with the noise generator off and $T_b$ is the data bit's duration.

The bit duration ($T_b$) in (5.2) needs further explanation. In a direct sequency spread spectrum modem, each bit is multiplied by the spreading code. Thus each bit consists of a large number of code chips which have chip durations ($T_c$) much shorter than the original data bit (see Figure 3.4). How-
ever, the chip duration and the bit duration are related to the spread spectrum processing gain ($G_p$).

$$G_p = \frac{T_b}{T_c} = \frac{f_c}{T_b}$$  \hfill (5.3)

Therefore, from (5.2) and (5.3) the energy per chip ($E_c$) is:

$$E_c = \left(\frac{v_{RMS}}{N}ight)^2/f_c = E_b/G_p$$  \hfill (5.4)

Equation (5.4) indicates that either the bit energy ($E_b$), or the chip energy ($E_c$) multiplied by the spread spectrum processing gain ($G_p$), can be used in equation (5.1).

The noise power spectral density is

$$N_o = \left(\frac{v_{N_{RMS}}}{N}ight)^2/f_R$$  \hfill (5.5)

where $v_{N_{RMS}}$ is the RMS voltage measured with the transmitter off and $f_R$ is the receiver's noise bandwidth. The receiver's noise bandwidth ($f_R$) is determined by the line coupling network. The noise bandwidth was measured by applying white noise to the receiver and observing the signal, after it has passed through the coupling network, on a spectrum analyzer. Figure 5.2 is a spectrum analyzer photograph of this signal and hence the receiver's noise bandwidth. As shown in Figure 5.2, the half power (3 dB) bandwidth is about 300 KHz.

Table 5.1 is a list of the data taken in the white noise environment. Figure 5.3 is a graph of this data along with the predicted performance as given by (5.1). Note that there is close agreement between the measured and predicted performance of the modem. This agreement indicates the modem is operating properly.
Vertical = 10 dB/cm.

Horizontal = 100 KHz/cm.

Figure 5.2 The receiver's noise bandwidth.
Figure 5.3 Predicted and measured performance.
Table 5.1

5.3 Performance in a Large Multi-Use Building

Figure 5.4 shows the equipment used for the tests which were carried out in the Hector MacLeod Building - a large multi-use building on the University of British Columbia campus. Synchronization is again provided by the power line. In this case, however, the receiver accepts not only the PLC signal generated by the transmitter but also the background noise which is present on the power line. In addition, both the transmitter and the receiver are connected across the power line impedance, which is both time and frequency dependent.

The measurements were taken over a 30-hour period on Sunday, November 11 and Monday, November 12, 1981. During this time the background noise level, as measured at the receiver, remained constant. There was, however, a period of two hours, between 2:00 p.m. and 4:00 p.m., Sunday afternoon, when many voltage transients occurred, presumably from loads switching on and off. During this period the error rates increased by more than a factor of 10.

Figure 5.5 is a plot of the data. Shown in Figure 5.5 are curves which show the best, worst, average, and white noise performances of the modem.
Figure 5.4 The equipment used for tests in a large multi-use building.
Figure 5.5 Performance in a large multi-use building.
5.4 Performance in a Small Residential Apartment Complex

The equipment used in this test was identical to that shown in Figure 5.4. The test was conducted in a small residential apartment complex, located in a residential district of Vancouver, B.C., containing approximately thirty apartments. The measurements were made over a 24-hour period on Tuesday, November 12, and Wednesday, November 13, 1981. In this case the background noise level varied by about 10 dB over the 24-hour period. The highest noise level occurred in the evening around 11:00 p.m. while the lowest noise level was observed in the early morning at about 4:00 a.m.

Figure 5.6 is a plot of the measurements taken. Included in Figure 5.6 are four curves showing the best, worst, average, and white noise performance of the modem.

5.5 Summary of Results

Observations made during the modem tests indicate that most of the errors occurred as a result of impulse noise as shown in Figure 2.9. Impulse noise presents a problem because the duration of the impulse is less than the duration of the code chips. For that reason, the impulses will pass almost unaffected through the data despreader as shown in Figure 5.7. After passing through the data despreader the impulses will cause step changes in the integrator's output which may cause it to cross over the comparator's threshold resulting in a bit error.

From Figures 5.5 and 5.6 it is apparent that there is a large variation in error performance for the same SNR, especially at low power levels. This is a result of the impulse noise. The impulses result from the switching of
Figure 5.6  Performance in a small apartment complex.
Figure 5.7  Effect of the data despreader on impulse noise.
electrical loads and therefore occur randomly. For a constant received
signal power the error performance will be largely determined, at least for
low signal power levels, by the number of impulses which occur during the
test.

Figures 5.5 and 5.6 also indicate that the error performance becomes
more consistent when the SNR is high. This improved consistency occurs
because the noise impulses will have less energy than the received signal.
At high power levels, the output of the integrator will be higher and the
step changes which result from the impulse noise will be less likely to cause
a decoding error. Instead, most errors will occur from the background noise
which is relatively constant.

It also became clear during the tests that the performance of the modem
varies widely over a 24-hour day, because background electrical interference,
which affects the modem's performance varies with the time of day. The modem
had the worst error performance in the evening, from 6:00 p.m. to 11:00 p.m.,
when electrical noise levels were highest. Conversely, the modem worked best
in the early morning hours from 3:00 a.m. to 6:00 a.m. when there was little
electrical noise. The difference between the best and worst performance over
two orders of magnitude for a constant SNR or, equivalently, over 5 dB in SNR
for the same error rate. When the test results are averaged the modem per­
forms as though it were in a white noise environment with a performance loss
of approximately 3 dB.

From Figures 5.5 and 5.6 a 12 dB SNR is required for an error rate of
better than one in 10,000 (10^-4). Using the peak noise levels observed
during tests the plus 12 dB SNR corresponds to a received signal power of
about one quarter of a milliwatt (-6 dBm). Therefore, one quarter of a milliwatt or more must be delivered to the receiver to obtain acceptable error performance. A typical PLC transmitter would deliver about 1 watt (+30 dBm) of power so that transmission lines losses between the transmitter and receiver must be kept below 36 dB for reliable communications. As a comparison, measurements made during the performance tests show an attenuation of 6 dB between the transmitter and receiver when they are both on the same side of the distribution transformer. The losses increased to 16 dB when the transmitter and receiver were on opposite sides of the transformer.

The results summarized in this section are based on measurements made on some typical power lines. A spread spectrum PLC modem should have a more consistent performance than a PLC modem using narrowband signalling. This is because many of the transmission impairments of the power system affect only a narrow frequency band and therefore will not significantly degrade the wideband SS signal. However, additional tests carried out on a wide variety of power lines will be necessary to fully assess the SS PLC modem's performance.
6. CONCLUSIONS

6.1 Summary

This thesis has documented the design and testing of a spread spectrum modem for communication over electrical power lines. The availability of a power line modem would allow many communications services within buildings, within cities, and between cities to be implemented inexpensively because the communications channel would be available without cost. In spite of its advantages, power line communications has not enjoyed widespread use. This is partially because there is not yet a consensus on the best signalling format for power line communications channels.

It is apparent from examining the transmission characteristics of power lines that many impairments affect only a narrow range of frequencies. For this reason a signalling format which uses a broad range of frequencies should perform better and more consistently than a signalling format which uses a narrow frequency range. Spread spectrum modulation provides such a broadband signalling format.

During the course of this work one other author [38] has suggested that spread spectrum communications should work well on power lines. However, no implementation details or performance results were published.

In order to test the performance of spread spectrum modulation on power lines a transmitter and receiver were designed and built. As a result of this design a novel synchronization procedure based on the 60 Hz power signal, was developed. This synchronization procedure significantly reduced the complexity of the modem. In addition, explicit relationships between
the modem's processing gain and the synchronization accuracy were established.

The modem was first tested in a controlled environment, in order that its actual performance could be compared to that theoretically predicted. The two results were very close which indicates that the synchronization procedure and other modem circuits operate properly. Further tests were carried out in a large multi-use building and a residential apartment complex. The results of these tests indicate that the modem performs well on actual power line circuits. The modem's performance varied substantially with the time of day as a result of a 10 dB variation in background noise levels during a twenty-four hour period. It was established that the modem requires a received signal to noise ratio of 12 dB to deliver a bit error rate below $10^{-4}$. Typical background noise levels and typical transmitted power indicate that the modem's performance will be adequate with channel losses as high as 36 dB.

6.2 Cost of the Spread Spectrum Power Line Modem

A spread spectrum modem is more complex and hence more costly than a narrowband modem. However, for power line communications the modem's synchronization can be based on the 60 Hz power signal. In this case complexity and cost are significantly reduced and only a small cost penalty is paid for the increased performance.

Because the implementation is not finalized, the modem's cost is difficult to estimate. Parts costs for the transmitter and receiver were about thirty dollars each. However, if power line communications becomes
widespread volume discounts and custom integrated circuits would significantly reduce these costs.

6.3 Suggestions for Future Work

Observations made during the modem tests indicate that impulse noise on the power line is a major source of bit errors, especially at low signal to noise ratios. Impulse noise is also a problem in other communications receivers such as FM demodulators. As a result impulse filters have been developed. Further work could investigate the effect of an impulse filter placed ahead of the spread spectrum receiver.

Phase noise on the 60 Hz power signal limits the maximum processing gain of the modem. Additional work on phase averaging the power signal could reduce this phase noise, improve the processing gain, and hence the modem's performance.

Although the transmission characteristics of the high voltage and distribution networks have been studied, very little work has been done on intrabuilding wiring. Future work could examine transmission characteristics of intrabuilding wiring especially at high frequencies.

Standard test procedures would be useful in comparing the performance of different PLC modems. Work could be done to develop a simulation of the transmission characteristics of the power system. This would allow modems to be tested in identical environments by operating on identical simulated power lines.
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