Chapter 3

Study and Modeling of Indoor Power Line Transfer Characteristics

3.1 INTRODUCTION

The transfer function of a channel defined by 
\[ H(f) = |H(f)| e^{\jmath \phi(f)} \]
and \( \phi(f) \) are the magnitude and phase response respectively is an important parameter for channel characterization. Many features of communication like channel capacity, modulation techniques and bandwidth allocation [41] are determined by its nature and variation. PL are found to be very harsh environment for data transfer suffering high attenuation [9] and variability with time, frequency and location [59] more than any other communication media available. Signal transmission over power lines suffer many losses some of which are as follows.

3.1.1 Coupling Loss. The couplers used to inject and receive the signal to/fro the PL introduces loss which can be minimized only if its output impedance matches the input impedance of the PL at the feeder point. The PL network behaves as a parallel resonant circuit with inductive behavior below the resonance frequency and capacitive impedance above it [22], [23]. At higher frequencies, the impedance is more resistive and has asymptotical behavior affected by its characteristic impedance. Some papers have reported the network impedance to vary between 20-190 \( \Omega \) [22] while others between a few ohms (~15\( \Omega \)) to k\( \Omega \) (~2k\( \Omega \)) [23] with a peak impedance between 4-20 MHz. The PL impedance therefore is a time and frequency varying quantity depending on the nature and distribution of the loads. The highly varying network impedance gives rise to severe impedance mismatch with the coupler and cause attenuation of the signals as high as 15dB [23].
3.1.2 Branching Loss. The PL network is constituted of many branched networks with different loaded condition which act as pieces of TL giving rise to impedance discontinuities. When a signal comes across a discontinuity, a part is transmitted and the rest is reflected causing echoing of the signal and occurrence of deep notches in the transfer characteristics [22]. The position of the notches are again time and frequency dependent. Loads with low impedances also exaggerate the notches and cause strong attenuation as high as 25dB [23]. These are dominant factors for indoor PLC.

3.1.3 Line Losses. The PL has got some impedance which is dependent on the type, length and size of the wire used. The attenuation of the signal usually increases with frequency because the lines are designed for transmission of low frequency 50 Hz signal only. These are the dominant factors mainly for outdoor long distance PLC. For a typical LV outdoor line, the attenuation increases with frequency and is approximately 15dB/km at 1 MHz and 50dB/km at 20 MHz [33].

3.1.4 Radiation Loss. At high frequencies, the length of the PL is comparable to the wavelength of the signal used for communication. The PL therefore acts as an antenna radiating [60] the signal into space. These emissions relate directly to the electrical unbalance of the PL network caused by the various network components which generate "common mode current" [60], [31] and the transmitted signal power [60]. The gain of the PL as a radiator increases rapidly with frequency. A radiating conductor with relatively low emissions at 0.1 MHz can have emissions tens of dB higher at HF. The radiation is one of the problems to be tackled for BPL.

Amongst the four types of losses, the coupling loss can be reduced by designing the coupler in a better way and the other two i.e. the branching and line loss cannot be controlled. The fourth loss can be decreased by decreasing the power level of the signal and introducing efficient communication schemes. Modeling the channel characteristics for selecting suitable modulation techniques, consistent with the regulatory constraints
with regard to EMC of the region [59] is therefore an important part of implementation of a suitable communication system. A summary of the available models is discussed in the following section.

### 3.2 REVIEW OF AVAILABLE POWER LINE CHANNEL MODELS

In order to simplify problems a lot of efforts are made to characterize and model the LV indoor PL channel, the reliability and accuracy of which is determined by two important factors- the model parameters and the model algorithms. Depending on the methodology of obtaining the model parameters, the available techniques can broadly be classified into the top-down and the bottom-up approach. In the top-down approach, requiring little computation, the model parameters like delay ($\tau$), attenuation and phase shift ($\phi$) are estimated only after the measurement of the actual TR and hence are prone to errors[22], [61], [62]. Besides, this methodology does not take into account the resonant effects due to parasitic capacitances and inductances and hence offers an incomplete description of the PL channel [63]. On the other hand, the bottom-up approach is free from these errors because the model parameters like the propagation constant ($\gamma$) and the characteristics impedance ($Z_0$) are derived theoretically [64], [65]. This is done either by using the analysis of network matrices as in MTL theory [64] or from the lumped circuit TL theory [65]. Though this method requires more computation, it is easy to predict changes in the TR should there be any change in the system configuration. In both the approaches, the modeling algorithms can be achieved in the time domain or the frequency domain. In the time domain modeling [22], [61], the PL is regarded as a multipath environment and an echo model is developed to represent its characteristics. This method is easier to implement in the top-down approach. Here the impulse response of the echo model $h(k)$ is written as a sum of delayed and weighted dirac pulses, needing $3*N$ parameters for a channel with $N$ paths, each paths needing the three model parameters as
stated earlier. In the bottom-up approach however the background reflections for impedance discontinuities are taken to be negligible [64]. In the frequency domain modeling, the network is regarded as a composition of many cascaded sections [64], [65], the overall TR written in terms of the transmission matrix or the scattering matrix of the cascaded portions. The frequency domain modeling is more friendly for complex multiple networks than the echo model.

The attempts to model the TC of PL vary world wide due to its dependency on the particular cabling and wiring practices [66]. In Singapore [67], the LV indoor PL cables are made of stranded copper conductors with PVC insulation placed inside a metal conduit that are embedded inside the concrete wall. In some European countries [33], the indoor distribution network cable consists of three wires with sector conductors covered by an outer conductor which functions as PEN (protective earth and neutral). After extensive study, we conclude that there is no universally accepted PL channel model. Also, there is a variation in the observed variables leading to different choice of modulation schemes. For example, the measured attenuation at a particular frequency as reported in different papers is not consistent and sometimes contradictory [4]. In terms of the phase distortion produced, some papers report that PSK modulation can be used while others report that FSK gives better performance [4]. In India, the indoor PL cabling wires are quite different consisting of two unshielded multi stranded wires (phase and neutral) including return ground wire individually covered by PVC insulation, usually enclosed in a plastic frame (~1" width), the distance between the wires being a variable term. A review of literature shows very little work done on this field in India. It is therefore necessary to carry out similar studies in our PL network in order to set design standards for systems in India.

In this chapter, we aim to study and model the TC of a typical Indian PL network and model the same for simulation in the subsequent chapters. Section 3.3 describes the
study of the TC of a floor in a university building with an estimate of the rate of attenuation with distance and frequency. Section 3.4-3.5 tests the applicability of a frequency domain two wire TL theory using the transmission matrix to an indoor PL setting. The more realistic MTL theory on the other hand, incorporating the third ground conductor and particular wiring techniques [12], involves many parameters and is difficult to determine with sufficient accuracy. The methodology used and the physics behind each setup is discussed in the contents. We conclude with some key findings and a proposal of a suitable communication system to overcome the odds.

3.3 STUDY OF TRANSFER CHARACTERISTICS OF A TYPICAL INDOOR POWER LINE SETTING

The TC of a typical indoor PL is studied in the first floor of the Department of Instrumentation and USIC, Gauhati University, Assam, India over a maximum wire length of 62m. The distribution network (Fig. 3.1) in the measurement site consists of two distribution boxes (DB-1 and DB-2), the AC power being fed to DB-1 from the distribution room (DR) situated in the ground floor of the building. DB-1 distributes power to various rooms. The loads are the usual laboratory and household equipments like computers, fan, bulbs and tubes. The attenuation is measured in the evenings when the variation of the loads in the building is minimal to minimize the variation of attenuation due to the time dependence of loads.
3.3.1 Experimental Setup. The experimental setup for measurement of the TC (Fig. 3.2) of a typical building consists of the following sections.

- **Signal generator.** The signal generator (IE 909, International Electronics) used is capable of generating frequencies up to 75 MHz.
- **Power amplifier.** The output signal from the signal generator has very low power and must be amplified to adequate levels before being transmitted over the PL. The circuit designed (APPENDIX A-4 & C-1) consists of a voltage amplifier and a power amplifier.
section. The voltage amplifier consists of two 2N3049 transistors capable of amplifying the low voltage from the signal generator to 1 Vrms. The power amplifier section consists of BD139 transistors that amplify the power of the signals (upto 15 MHz) to 1 watt. The transformers of the power amplifier are adjusted to adequately amplify the signals in the bandwidth of observation.

3.3.1.3 The couplers. The transmitter and receiver are connected to the PL through adequate couplers acting as HPF having cut off frequency of 3 kHz to cut off the 230V, 50 Hz AC mains from harming the connected instruments (Fig. 2.2 and Table 2.1 of Chapter 1).

3.3.1.4 The measuring instrument. The receiver coupler is connected to the \( \text{which is placed in FFT mode to measure the amplitude of the received signal.} \)

3.3.2 Methodology. Signal at 4.5 Vrms (13dB) is injected to the PL via plug points and the attenuated signal is received at various plug points in various rooms. The input signal is varied in steps of 5 kHz in the frequency range 10 kHz to 1 MHz and in steps of 100 kHz from 1 MHz to 15 MHz. The signal is transmitted both in the phase-ground (P-G) and phase-neutral (P-N) pairs. The magnitude of the experimental transfer TR is given by

\[
|H(f)|_{\text{exp}} = 20 \log_{10} \frac{\text{Received Signal Amplitude}}{\text{Transmitted Signal Amplitude}} dB
\]

3.3.3 Experimental Results. In all the cases studied, the indoor PL showed a low pass characteristics (Fig. 3.3(a-b) and 3.4) with the maximum usable frequency (MUF, the maximum frequency where the received signal strength meets the noise floor) decreasing with increasing length (Table 3.1) of the wires. The received signal strength or the TR is not only dependent on the length of the wire but also on the frequency of the transmitting signal. In macroscopic terms, there is increase in attenuation with frequency and also with distance between the source and the transmitter (Fig. 3.5) as given by the two parameter model (Parameters in Table 3.1):
20\log|H(f)| = -(\alpha d f + \beta) .................................................. 3.2

Fig. 3.3: The received signal amplitude in the frequency range (a) \( \leq 1\text{MHz} \) (P-N) and (b) \( 1\text{MHz} \leq f \leq 15\text{MHz} \) (P-N)

Fig. 3.4: The received signal amplitude in the frequency range 100 kHz -15 MHz for P-G.
The received signal or the TR is characterized with significant notches with the transfer function falling as low as -60dB at these frequencies. Some of these notches are stagnant with length (for example at 300 kHz) while others vary with different setups proving them to be dependent on the varying load condition. On an average there is an attenuation of 0.028 dB per unit length per unit frequency for the setups with loaded condition (6 to 2, 3 & 4). Significant changes are also seen in the TR of loaded conditions (6 to 2, 3 & 4) as compared to the cable with a direct line (6 to D). For example the attenuation of the received signal is large compared to a wire of same length without loads. The received signal in the path 6 to 2 is nearly 22dB less than the path 6 to D though the difference in the length in the paths is only two meters. This also proves the dependency on the load conditions more than the attenuation of the cables. The MUF in the P-N pair (Fig. 3.3(b)) is found to be much smaller than that in the P-G pairs (Fig. 3.4) (~ by 2 MHz). This is because most of the loads are connected to the PL in the P-N pair.

Table 3.1: Transfer characteristics model parameters

<table>
<thead>
<tr>
<th>Setup</th>
<th>d</th>
<th>α</th>
<th>β</th>
<th>$R^2$</th>
<th>MUF(P-N)</th>
<th>MUF(P-G)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6 to D</td>
<td>30m</td>
<td>0.0402</td>
<td>3.471</td>
<td>0.6703</td>
<td>&gt;15</td>
<td>-</td>
</tr>
<tr>
<td>6 to 2</td>
<td>32m</td>
<td>0.0335</td>
<td>45.561</td>
<td>0.5849</td>
<td>13</td>
<td>-</td>
</tr>
<tr>
<td>6 to 3</td>
<td>34m</td>
<td>0.0832</td>
<td>43.588</td>
<td>0.5233</td>
<td>7.2</td>
<td>9</td>
</tr>
<tr>
<td>6 to 4</td>
<td>37m</td>
<td>0.1745</td>
<td>35.713</td>
<td>0.6745</td>
<td>3.2</td>
<td>5.2</td>
</tr>
</tbody>
</table>
3.4 MODELING OF THE TRANSFER FUNCTION

3.4.1 The Transmission Line Theory. The TL theory describes the propagation of an EM wave in a TL. The model assumes that the signal injected into the TL propagates along the line as a TEM wave. When the length of the cable is short compared to the wavelength of the EM wave, the TL analysis is not necessary (short when the electrical length is less than \( l < \lambda/16 \) [68] or \( l < \lambda/8 \) [69] where \( \lambda \) is the wavelength of the signal used) instead may be modeled by using a simple model consisting of its distributed components. The PL can be considered as a two wire TL with distributed parameters R, L, G and C as shown in Fig. 3.6.

Fig. 3.6: The transmission line.

The quantities \( v(x,t) \) and \( v(x + \Delta x,t) \) denote the instantaneous voltages at location \( x \) and \( x + \Delta x \) respectively. In the same way, \( i(x,t) \) and \( i(x + \Delta t) \) denote the instantaneous currents at \( x \) and \( x + \Delta x \) respectively. R defines the resistance per unit length for both the conductors (\( \Omega/m \)), L is the inductance per unit length (\( H/m \)), G is the conductance per unit length (\( mho/m \)) and C is the capacitance per unit length (\( F/m \)). Simplification of Kirchoff's voltage and current law gives,
\[
\frac{d^2V(x)}{dx^2} = \gamma^2 V(x) \\
\frac{d^2I(x)}{dx^2} = \gamma^2 I(x)
\]

where

\[
\gamma(f) = \sqrt{(R + j\omega L)(G + j\omega C)} = \alpha(f) + j\beta(f)
\]

\(\gamma(f), \alpha(f)\) and \(\beta(f)\) are known as the propagation constant, the attenuation constant (in Np/m) and phase constant (in rad/m) respectively.

The voltages and current at \(x\) can also be expressed as

\[
V(x) = V^+ (x) + V^- (x) = V_0^+ e^{-\gamma x} + V_0^- e^{\gamma x}
\]
\[
I(x) = I^+ (x) + I^- (x) = I_0^+ e^{-\gamma x} + I_0^- e^{\gamma x}
\]

where the plus and minus subscripts denote waves traveling in the forward and backward direction respectively. The characteristic impedance is defined by

\[
Z_0 = \frac{V^+}{I^+} = \frac{V^-}{I^-} = \sqrt{\frac{R + j\omega L}{G + j\omega C}}
\]

The line parameters \(R, L, G, C\) of the parallel wire TL is calculated using the formulae

\[
R = \frac{1}{\pi \mu_0 \sigma_{\text{cond}} D}; \quad L = \frac{\mu_0}{\pi} \text{acosh} \left( \frac{D}{2a} \right); \quad G = \frac{\pi \sigma_{\text{die}}} {\pi \text{acosh} \left( \frac{D}{2a} \right)}; \quad C = \frac{\pi \varepsilon} {\pi \text{acosh} \left( \frac{D}{2a} \right)}
\]

where

\(a\) : radius of the conductor
\(D\) : distance between the cables
\(\mu_0\) : permeability of the conducting medium
\(\sigma_{\text{cond}}\) : conductivity of the conducting Medium
\(\sigma_{\text{die}}\) : conductivity of the dielectric medium between conductors
\(\varepsilon\) (=\(\varepsilon_0\varepsilon_r\)) : permittivity of dielectric material between the conductors.
Eq. 3.6 cannot however be applied to the PL wires directly because of its dependency on frequency, nature and type of wiring used as stated in Sec 3.4.2.

3.4.2 Correction in Line Parameters [70]-[73].

3.4.2.1 Correction in resistance.

3.4.2.1.1 Correction at high frequency. When an AC flows in a conductor, the self inductance of the conductor causes the current to be confined at the surface of the wire (skin effect). As the frequency \( f \) of the AC increases, the resistance of the wire also increases [71]. The skin depth of the wire (the distance through which the amplitude of the traveling wave decreases by a factor of \( e^{-1} \)) is given

\[
\delta = \frac{1}{\sqrt{\pi f \mu \sigma_{\text{cond}}}} \tag{3.7.a}
\]

The resistance of the conductor of radius \( a \) is therefore given by

\[
R = \frac{1}{\pi a \delta} \sigma_{\text{cond}} \tag{3.7.b}
\]

3.4.2.1.2 Correction for multi-stranded nature. When the conductor is multi-stranded [73], the resistance again increases because the area of current flow is again reduced. This is because there are gaps at the circumference of the conductor wire. The ratio of the effective area to the total area is given by

\[
X_r = \frac{\text{Effective area}}{\text{Total area}}
\]

\[
= \frac{ACos \left[ \frac{r-\delta}{r} \right] \cdot r^2 - (r-\delta) \sqrt{r^2 - (r-\delta)^2}}{2 \cdot r \cdot \delta}
\]

where \( r \) is the radius of a single wire in the stranded conductor and \( \delta \) is the skin depth given by Eq. 3.7.a. The corrected resistance is then given

\[
R_{\text{corrected}} = X_r R \tag{3.8}
\]
3.4.2.2 Corrections in capacitance. The high frequency communication signals in PL are not only coupled between the two pairs of lines (Fig. 3.7), but also coupled from the live cable to the earth cable first and then coupled from the earth cable to the neutral cable [32]. This adds a term $C_{cable}/2$ to the capacitance term.

![Diagram of capacitance](image)

Fig. 3.7: Corrections in capacitance.

The total cable capacitance is given by

$$C_{total} = C_{cable} + \frac{C_{cable}}{2}$$

Where $C_{cable}$ is the capacitance per unit length given by Eq. 3.6.

3.4.2.3 Corrections in conductance. The conductance has the same corrections factor as the capacitance. The total conductance is

$$G_{total} = G_{cable} \cdot \frac{C_{total}}{C_{cable}}$$

3.4.2.4 Corrections in inductance. The inductance of the two wires TL includes the self inductance of each conductor and the mutual inductance between them. The mutual inductance between a pair of parallel conductors [32] is given by

$$I_m = \frac{\mu_0}{\pi} \ln\left(\frac{D-a}{a}\right) \text{H/m}$$

Where $D$ is the distance between the two conductors and $a$ is the radius of the conductor.

The total inductance is written as

$$L_{total} = L + L_m$$

Here $L$ is the inductance per unit length given by Eq. 3.6.
3.4.3 The ABCD Matrix. The TL can also be considered as a cascade of a number of two port devices each unit defined by their ABCD or the TL matrix [70]. The chain or ABCD matrix [70] of a two port system (Fig. 3.8) is defined as

\[
\begin{bmatrix}
v_1 \\
i_1 \\
v_2 \\
i_2
\end{bmatrix} =
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}
\begin{bmatrix}
v_2 \\
i_2
\end{bmatrix}
\]

Where

\[
A = \frac{v_1}{v_2} |_{i_2=0} \quad B = \frac{v_1}{-i_2} |_{v_2=0} \quad C = \frac{i_1}{v_2} |_{i_2=0} \quad D = \frac{i_1}{-i_2} |_{v_2=0}
\]

The ABCD -parameter is most suitable when cascading networks. The overall ABCD matrix of two cascaded networks with individual networks with parameters \(A'B'C'D'\) and \(A''B''C''D''\) is given by the product of the ABCD matrices of individual sections i.e.

\[
\begin{bmatrix}
v_1 \\
i_1 \\
v_2 \\
i_2
\end{bmatrix} =
\begin{bmatrix}
A' & B' \\
C' & D'
\end{bmatrix}
\begin{bmatrix}
A' & B' \\
C' & D'
\end{bmatrix}
\begin{bmatrix}
v_2 \\
i_2
\end{bmatrix}
\]

In a similar way a number of ABCD matrices can be multiplied to get the overall transmission matrix of the entire network.

![Fig 3.8: Two port network.](image)

3.4.3.1 The ABCD matrix of some useful configurations [70], [71]. The ABCD matrices of some useful configurations are given in Fig. 3.9.
In the Fig. 3.9, $Z_m$ is the input impedance of the branched network given by

$$Z_m = \left( Z_o + Z_{br} \tanh(\gamma_{br} d_{br}) \right) / \left( Z_o + Z_{br} \tanh(\gamma_{br} d_{br}) \right)$$

where $Z_o, \gamma_{br}, Z_{br}, d_{br}$ are the characteristics impedance and propagation constant of the branched circuit, the load and the length of the branch respectively.

The ABCD matrix helps in the determination of some useful circuit parameters like the input impedance and the transfer function.
3.4.3.2 Input impedance of a 2PN. The input impedance of a 2PN in terms of the ABCD matrix is given by

\[
Z_n(f) = \frac{A(f)Z_{br}(f) + B(f)}{C(f)Z_{br}(f) + D(f)}
\]

where \( Z_{br}(f) \) is the value of the load at the terminal of the branch. If however, the branching connection is un-terminated, the input impedance is given by

\[
\lim_{Z_{br} \to \infty} Z_n(f) = \lim_{Z_{br} \to \infty} \frac{A(f)Z_{br}(f) + B(f)}{C(f)Z_{br}(f) + D(f)} = \frac{A(f)}{C(f)}
\]

3.4.3.3 Transfer function of a network using ABCD matrix. The transfer function of a network in terms of the ABCD matrix is given by

\[
T(f) = \frac{V_L(f)}{V_S(f)} = \frac{Z_L(f)}{A(f)Z_L + B(f) + C(f)Z_LZ_L + D(f)Z_L}
\]

Here, \( V_L = V_2 \) as in Fig. 3.9.

3.5 METHODOLOGY USED TO TEST THE TRANSFER FUNCTION

To test the TL theory in PL, the experimental TR of a PL network is compared with the theoretically obtained one.

3.5.1 The Experimental Transfer Function. The experimental network (Fig. 3.10) in which the theory is to be tested consists of a number of branches and loads connected using the typical wire (Specification as in Table 3.2 and APPENDIX B-2) used in general.

![Fig. 3.10: The power line network for testing TL theory](image)

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indoor household wiring. The network has a maximum wire length of 34 m. The point at PLI (power line inlet) connects the experimental network to the actual PL of the building. For different setups, (Table 3.2, Table 3.3) high frequency signals ($V_{rms} = 4.5$ V measured at the primary of the coupler in the signal generator side) from a signal generator through an adequate power amplifier are injected at PLUG 1 from 100 kHz -15 MHz in steps of 100 kHz through a coupler between the phase and neutral lines (Fig. 3.2). The received signal is observed at PLUG 7 with the help of a DSO in FFT mode through a similar coupler across a 50 ohm load both for PLI ON and OFF conditions. All readings are taken in the evening when there is slight change in the load conditions in the electrical network of the building. The TR is obtained experimentally given by Eq. 3.1.

3.5.2 The Theoretical Transfer Function. To obtain the theoretical TR, the line parameters (Eq. 3.6) are obtained for the particular cables used in indoor PL after applying the necessary corrections (Eq. 3.7-3.11) (Table 3.2). For house power cables, the dielectric material between cable conductors is inhomogeneous in both space and contents because the shape of the conductor and insulation is not perfectly round and the spacing is a mixture of insulation and air. In this work, the dielectric is considered to be a mixture of air and PVC and the shape of the wire round to keep the model tractable. The variation of the magnitude and phase of the characteristic impedance, attenuation constant and the phase constant with frequency (Eq. 3.4 and Eq 3.5) are shown in Fig. 3.11. For every line condition, the ABCD matrix of the entire network is calculated considering the network as a series combination of various network sections (Sec 3.4.3) as necessary (using MATLAB-7) and the theoretical TR ($\tau(f)$) of the network calculated (Eq. 3.14).
Fig. 3.11: Plot of the variation of the magnitude and phase of the $Z_o(f)$, $\alpha(f)$ and $\beta(f)$ with frequency.

However, the theoretical TR cannot be compared against the experimental one $|H(f)|_{\text{exp}}$ without the addition of a correction. The corrected theoretical TR i.e. $|H(f)|_{\text{theory}}$ is given by the magnitude of $H(f)$ where

$$H(f) = \frac{V_i(f)}{V_s(f)} = \frac{V_L(f)}{V_s(f)} \cdot \frac{V_s(f)}{V_i(f)} \quad (V_L(f) = V_2(f))$$

$$= T(f) \cdot \frac{V_s(f)}{V_i(f)}$$

$$= T(f) \cdot \frac{Z_m(f)}{Z_m(f) + Z_i} \quad \text{3.16}$$

Here, $Z_m(f)$ is the input impedance of the two-port network [20] (Eq. 3.12). For each setup, the experimentally obtained transfer function (Eq. 3.1) is compared with the magnitude of theoretically obtained one (Eq. 3.16).
Table 3.2: Specifications of the wiring, loads and components used to test applicability of TL theory to PL.

<table>
<thead>
<tr>
<th>Line Parameters</th>
<th>Coupler</th>
<th>Circuit Specification</th>
<th>Load</th>
</tr>
</thead>
<tbody>
<tr>
<td>R variable C1</td>
<td>0.1μF</td>
<td>Z₀ = 50Ω</td>
<td>Capacitor 1/jωC (C = 0.1μF)</td>
</tr>
<tr>
<td>L 0.69μH/m C2</td>
<td>0.1μF</td>
<td>Zₐ = 50Ω</td>
<td>Heater R + jωL (R = 48.5Ω, L = 103x10⁻⁴H)</td>
</tr>
<tr>
<td>G 0.018μmho/m L</td>
<td>1.5mH</td>
<td></td>
<td>Bulb 806Ω</td>
</tr>
<tr>
<td>C 33pF f_c</td>
<td>25 kHz</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

3.5.3 Case by Case Study. The magnitude of the derived and the experimental TR showed peaks and dips with an overall increasing attenuation at higher frequencies with considerable similarities. When an EM travels in a medium terminated by a low, high or equal to its characteristics impedance it undergoes reflection or complete absorption at the terminal depending on whether the terminated impedance is not equal or equal to the characteristics impedance. When the terminating line is short circuit, there is a voltage node and current antinodes at the terminating end. In an open circuit [72] line, the condition is vice-versa. When the line is terminated by impedance equal to the characteristics impedance of the TL then the voltage is uniform and no signal is reflected but entirely absorbed by the load.

3.5.3.1 Direct line with no loads (case 1). In the simplest cases tested (Fig. 3.12), the signal generators and the receiver are directly connected to the TL with no loads and additional branched lines through the couplers.

![Diagram](Fig. 3.12: The direct line with no loads and open circuits (Case 1)

In the Fig. 3.12, c₁ and c₂ are the capacitors used to couple the signal to and from the PL.)
It is observed from Fig. 3.13 that the experimental TR showed valleys and peaks at specific frequency intervals (2.2, 3, 4.8, 6.4, 8 and 9.6 MHz).

Table 3.3 The setups used to test the TL theory

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Case 1: direct line without loads</th>
<th>Case 2: open wire &amp; direct line</th>
<th>Case 3: capacitive load</th>
<th>Case 4: inductive load</th>
<th>Case 5: three loads</th>
</tr>
</thead>
<tbody>
<tr>
<td>L1(m)</td>
<td>34 m</td>
<td>6.65 m</td>
<td>6.65 m</td>
<td>6.65 m</td>
<td>6.65 m</td>
</tr>
<tr>
<td>L2(m)</td>
<td>27.35 m</td>
<td>1.25 m</td>
<td>1.25 m</td>
<td>1.25 m</td>
<td>1.25 m</td>
</tr>
<tr>
<td>L3(m)</td>
<td>0</td>
<td>26.1 m</td>
<td>26.1 m</td>
<td>6.73 m</td>
<td></td>
</tr>
<tr>
<td>L4(m)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>4.45 m</td>
<td></td>
</tr>
<tr>
<td>L5(m)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>14.92 m</td>
<td></td>
</tr>
<tr>
<td>L6(m)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td></td>
</tr>
<tr>
<td>L7(m)</td>
<td>0</td>
<td>6 m</td>
<td>2.55 m</td>
<td>2.55 m</td>
<td>2.55 m</td>
</tr>
<tr>
<td>L8(m)</td>
<td>0</td>
<td>0</td>
<td>2m</td>
<td>2 m</td>
<td>2 m</td>
</tr>
<tr>
<td>L9(m)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>2 m</td>
</tr>
<tr>
<td>L10(m)</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Z1(Load)</td>
<td>open</td>
<td>open</td>
<td>open</td>
<td>open</td>
<td>open</td>
</tr>
<tr>
<td>Z2(Load)</td>
<td>-</td>
<td>capacitor</td>
<td>heater</td>
<td>capacitor</td>
<td></td>
</tr>
<tr>
<td>Z3(Load)</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>bulb</td>
<td></td>
</tr>
<tr>
<td>Z4(Load)</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>heater</td>
</tr>
</tbody>
</table>

The variation showed considerable similarities with the derived TR obtained from the ABCD matrix of setup 1 given by

\[
Y = \begin{pmatrix}
1 & 0 \\
0 & 1
\end{pmatrix}
\begin{pmatrix}
1 & 1/jwc \\
jwc & 1
\end{pmatrix}
\begin{pmatrix}
Cosh(b1) & Z_0 \text{Sinh}(b1) \\
\text{Sinh}(b1)/Z_0 & Cosh(b1)
\end{pmatrix}
\begin{pmatrix}
1 & 1/jwc2 \\
jwc2 & 1
\end{pmatrix}
\begin{pmatrix}
1 & 0 \\
0 & 1
\end{pmatrix}
\]

Where \( b1 \) is the propagation constant of the TL of length \( L1 \). Here the frequencies of valleys and peaks (1.55, 3.08, 4.64, 6.26, ....... in MHz) are odd and even multiples of the fundamental (1.55 MHz) respectively with a velocity factor 0.66. The first dip (i) of Fig. 3.13 therefore corresponds to the half wave standing wave pattern on a \( \lambda/4 \) shorted TL (length 34 m). The consecutive dips at (iii) and (v) are the 3/2 and 5/2 standing wave pattern on a 1/4 shorted TL (length 34 m).
pattern on a $3/4$ and $5/4$ wave shorted TL respectively. Similarly, the peaks at (ii), (iv) and (vi) are the full wave, double wave and triple wave standing wave pattern on a half, full and $3/2$ wave shorted TL. The TR shows that the system is weakly resonant. The line terminated by the 50 ohm resistance acts as nearly short for the TL with characteristics impedance ~145 ohms.

![Graph showing TR of Case 1](image)

Fig 3.13: TR of Case 1

3.5.3.2 Direct line with an open circuit (case 2): The experiment and simulation is again performed for a network which consisted of an open circuit branch which is typical in indoor power lines (Fig. 3.14 & 3.15).

![Diagram of direct line with an open circuit](image)

Fig. .3.14: Direct line with an open circuit (Case 2)

In this setup, the ABCD matrix is obtained by
Y =

\[
\begin{pmatrix}
1 & 0 & 1 & 1/jw1 \\
0 & 1 & 0 & 1 \\
1 & 1/jw2 & 1 & 1/jw2 \\
0 & 1 & 0 & 1
\end{pmatrix}
\begin{pmatrix}
Cosh(b1) & Z_oSinh(b1) \\
Sinh(b1)/Z_o & Cosh(b1)
\end{pmatrix}
\begin{pmatrix}
1 & 0 & 1 & 1/jw3 \\
0 & 1 & 0 & 1 \\
1 & 1/jw3 & 1 & 1/jw3 \\
0 & 1 & 0 & 1
\end{pmatrix}
\begin{pmatrix}
Cosh(b2) & Z_oSinh(b2) \\
Sinh(b2)/Z_o & Cosh(b2)
\end{pmatrix}
\]

\[zlcrl = \frac{Cosh(b3)}{Sinh(b3)} \frac{1}{Z_o}\]

Where, \(zlcrl\) is the input impedance at the input of the open circuited line of length \(l1\) and \(b1, b2, b3\) are the propagation constant of the TL sections of length \(L1, L2\) and \(l1\) respectively. Both showed a drastic change compared to case 1 with notches in the TR over and above the characteristics of weakly resonant circuit. When an EM wave traveling through a two wire TL meets an open circuit, the wave is completely reflected at the terminal with voltage maxima at the terminating end. This wave again travels back to the incident path. The received signal at the terminating end therefore consists of a direct signal from the transmitter and a reflected signal from the open circuit end. If the phase difference between the two signals is such that they completely nullify each other then a deep notch is found at the particular frequency. In the circuit used, a direct wave reaches the load in the path A→B→C and a reflected wave in the path A→B→D→B→C. In the Fig. 3.15, the path difference between the direct (path length 34m) and the reflected (path length 46m) wave at the open circuit is \(\sim \lambda/2\) at 8.7 MHz and hence the deep notch. More notches can therefore be predicted at \(n\lambda/2\) where \(n=2, 3, 5 \ldots \) etc. The position of the notches is highly dependent on the length of the open circuited wire and the distance between the transmitter and the receiver.
3.5.3.3 Line with an open circuit and capacitive/inductive load (case 3 and 4). Here we test an open circuit and a reactive load like a 0.1 μF capacitive (case 3) or inductive load (case 4) (Fig. 3.16). The capacitive load is general in computers and television sets to meet electromagnetic emission regulations [24]. The inductive load is found in household heater.

![Fig 3.16: Line with open circuit and capacitive/inductive load (Case 3 and Case 4).](image)

The ABCD matrix of the network is given by

\[
Y = \begin{bmatrix}
1 & 0 & 1/jw & 1 \\
0 & 1 & 1 & 0 \\
1/jw & 1 & Cosh(b1) & Z_o Sinh(b1) \\
1 & 0 & Z_o Sinh(b1)/Cosh(b1) & 1
\end{bmatrix}
\begin{bmatrix}
1 & 0 & 1/jw & 1 \\
0 & 1 & 1 & 0 \\
1/jw & 1 & Cosh(b2) & Z_o Sinh(b2) \\
1 & 0 & Z_o Sinh(b2)/Cosh(b2) & 1
\end{bmatrix}
\begin{bmatrix}
1 & 0 & 1/jw & 1 \\
0 & 1 & 1 & 0 \\
1/jw & 1 & Cosh(b3) & Z_o Sinh(b3) \\
1 & 0 & Z_o Sinh(b3)/Cosh(b3) & 1
\end{bmatrix}
\begin{bmatrix}
1 & 0 & 1/jw & 1 \\
0 & 1 & 0 & 1
\end{bmatrix}
\]

\[
zclr1 = \frac{Cosh(b4)}{Sinh(b4)}
\]
\[ z_{lc}r_2 = Z_o \frac{Z_2 + Z_o \tanh(b5)}{Z_o + Z_2 \tanh(b5)} \]

Where \( z_{lc}r_1 \) and \( z_{lc}r_2 \) are the input impedance of an open circuited line of length \( l_1 \) and impedance as seen at the input of the second branch of length \( l_2 \) terminated by an impedance \( Z_2 \). \( b1, b2, b3, b4 \) and \( b5 \) being the propagation constants of the branches of length \( L1, L2, L3, l_1 \) and \( l_2 \) respectively.

For case 3, the TR shows a deep notch at 400 kHz (viii) of Fig. 3.18. The capacitor acts as an open circuit at low frequencies and a short circuit at high frequencies. However the impedance applied at the TL is not the impedance of the capacitor, but the impedance seen at the input of the branch line 2 i.e. \( z_{lc}r_2 \) (Fig. 3.17) which shows a deep notch at 400 kHz. This can be explained by the fact that the absolute value of \( z_{lc}r_2 \) also shows a minimum value at this frequency. The line in both the cases acts as a weakly resonant line showing similar characteristics as those discussed earlier. For inductive load (case 4), the impedance increases with frequency and the line acts as an open circuit, therefore the transfer function is similar to that of a weakly resonant line (Fig. 3.18).

Fig. 3.17: Magnitude of the impedance of the capacitor and \( z_{lc}r_2 \).
3.5.3.4 Line with an open circuit, resistive, capacitive and inductive loads (case 5). Here we test a circuit with a capacitive, resistive and inductive load (Fig. 3.19). The resistive load is offered by a bulb with a resistance 80Ω at incandescent state.

The net ABCD matrix is the cascade of the individual ABCD matrices given by

\[
Y = \begin{bmatrix}
1 & 0 & \frac{1}{j\omega c_1} & Cosh(b_1) & Z_oSinh(b_1) \\
0 & 1 & 0 & \frac{1}{zlc_1} & Cosh(b_1) \frac{1}{Sinh(b_1)/Z_o} \\
\frac{1}{zlc_2} & 0 & Cosh(b_2) & Z_oSinh(b_2) \\
\frac{1}{zlc_3} & 0 & Cosh(b_3) & Z_oSinh(b_3) \\
\frac{1}{zlc_4} & 0 & Cosh(b_4) & Z_oSinh(b_4) \\
0 & 1 & 0 & 0 & 1 \\
\end{bmatrix}
\]

Where zlc_1, zlc_2, zlc_3 and zlc_4 are the input impedance of the open circuited line of length l_1, and line of lengths l_2, l_3 and l_4 terminated with loads Z_2, Z_3 and Z_4 respectively.

Here b_1, b_2, b_3, b_4 and b_5 being the propagation constants of the branches of length L1,
As seen from the Fig. 3.20, the extent of correlation between the simulated and the experimental line PLI-OFF decreases though there is some amount of similarity between the same. This is because in the PLI-OFF state, the resistance of the bulb is less than that which is considered in simulation and PLI-ON state. The additional peak (x) at 3 MHz (seen in PLI-ON state) can only be explained by the increased resistance of the bulb with temperature. It is also noted that the short circuit characteristics (ix) of \( zclrl \) maintains its behavior and overrules other effects.

### 3.5.4 Results of the Study.

From the above, it is observed that the TR of the PL can be simulated using the TL theory. The magnitude of the derived and experimental TR showed peaks and dips with considerable similarities although the attenuation of the later is more than the former due to improper wiring and attenuation of the high frequency signals in the wire used. The fact that the distance between the two wires is not a constant may also lead to such discrepancies. When the point PLI is made ON (connected to the rest of the PL network) further discrepancies are added although keeping some of the general trends. Additional notches are added (for example at 4.4 MHz) common to all conditions probably due to the addition of unknown loads in the network. Though an exact representation is not obtained, still the suitable transmitting frequencies for PLC can be predicted if the loads are known \textit{apriori}. The TL theory can be used as a channel simulator to replicate the PL channel to test various communication systems (Chapter 5). Slight
change in load condition will lead to considerable changes in the TR as seen from the simulation graphs in Fig. 3.21 when the capacitive loads is connected to the PLUG 3 and 6, with the position of the transmitter and the receiver unchanged in Section 3.5.3.3. A MATLAB simulation of the TR of a typical indoor PL (till 500 kHz) with four random loads shows considerable (Fig. 3.22) changes with unpredictable dips and notches (as in 0.125MHz of TR3) at different frequencies.

![Graphs showing TR changes](image)

**Fig 3.21:** Dependence of TR on the position of the capacitor.  
**Fig. 3.22:** Varying TR for four random loads.

### 3.6 SUMMARY

In this chapter, studies on the TR of a typical indoor PL setting are presented. The methodology has been used to simulate a PL channel described in Chapter 5 to test the efficiency of a suitable communication system. Estimation of a local PL channel capacity is done in the following chapter.