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Measurement and Characterisation of Impulse Noise on DSL for ISDN Applications

**by
Vivek Kapil, B. Eng.**

**A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfilment of the requirements
for the degree of
*Master of Engineering***

**Ottawa-Carleton Institute for Electrical Engineering
Faculty of Engineering
Department of Systems and Computer Engineering
Carleton University
August, 1989
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Acceptance

The undersigned hereby recommends to the Faculty of
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submitted by Vivek Kapil, B. Eng.

in partial fulfilment of the requirements for
the degree of Master of Engineering

Dr. David D. Falconer, Thesis Co-supervisor

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Carleton University, August, 1989

To my Parents and Gurus

Abstract

To meet the growing demand of the evolving Integrated Services Digital Network (ISDN), the present subscriber loop-plant would have to support data transmission for basic and primary access bit-rates (160kbps through 1.544Mbps). This loop thus becomes the critical channel of communication between the central office and the subscribers. Major networks and operating telephone companies have tried to study their loop-plants in order to characterise the physical and electrical characteristics. Impulse Noise was found to be a serious disturbance for bandlimited and power-restricted ISDN data transmission application.

A need has arisen to assess the high bit-rate performance of the existing loop-plant in an impulse noise environment. The theme of this research is to characterise the real-time measured impulse noise data in a business PBX¹ loop-plant environment and then assess the performance of the digital subscriber loops in this environment.

The measurement equipment design and the measurements were carried out for the initial period of this research. The impulse noise data was re-constructed and analysed for time domain and frequency domain characteristics. The amplitude and duration of impulsive events probabilities are presented with respect to the receiver threshold. Their arrival time statistics has also been displayed.

Finally, a complete 2B1Q digital subscriber loop system was simulated for conditional distribution of number of errors evaluation whenever there is an impulse on the loop. The DSL system performance was studied with an adaptive decision feedback equaliser (DFE) receiver only and a $(1-\eta D)$ filter + DFE receiver. The performance is found to be better with the latter receiver structure.

¹ Private Branch Exchange.

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Chapter 1

Introduction to Impulse Noise

1.1 General

Telephone jacks, those unobtrusive socket plaques screwed on the walls of every home and office in the civilised world, have recently become the key to the dream of integrated global communications.

The present day demand for a variety of new data services like simultaneous transmission of digitized speech and data (enhanced telephony), database communication capabilities, ASCII support (packet & circuit-switching), WAN², Shared screen, call management, etc. has guided the existing trend towards the integration of voice and data transmission networks. The existing analog telephone network is evolving towards the Integrated Services Digital Network (ISDN). Whether this dream of integrated global communication will come true or not, mainly depends upon the complete digitisation of all the telephony systems. By looking at the presently growing networks, we find that most of them are based on digital facilities for inter-office and inter-cities switching. So far most of the work has been done for digitisation of the networks, for example, digital switching has already become a reality, but the subscriber loop, a medium for

² Wide Area Network.

transmission, has still remained largely untouched by the digital trends. Even today, the final reach to the customers happens to be mostly analog! The various schemes of digital transmission over the existing loop plants are keys to the user-to-user connectivity in the realization of integrated services digital networks (ISDNs). This is obvious, because a large portion of the investment in the present telephone system is in the subscriber loop plant, which consists of twisted copper pairs that provide connection between the central-office (CO) and the subscriber. In the initial stages, the basic access to the ISDNs is planned to be provided over the two-wire loop plant.

It is a tantalizing prospect for telecommunications carriers that the ISDN promises to be able to provide an abundance of digital data services through a single technology, in common with the voice network. Conceptually, the ISDN is a uniform access system. It promises to provide all-digital, higher speed and standardized services with more sophistication, and to bring varying demands under the same umbrella. The ISDN also promises flexibility in providing appropriate features and services to the existing and new users. Thus leading to more economical equipment for both the carrier and the user.

In pre-ISDN services, in some countries, the basic access to the customers has been provided by integrated digital access at 80kbps by two-wire transmission over existing local network cable pairs[1]. Eventually, high network penetration is envisaged, and the basic access rate will be raised to the CCITT recommended rate of 144kbps with full duplex bit-stream. This data-rate includes provisions for two voice/data channels at 64kbps each and one data channel at 16kbps. These channels are shown in the Figure 1.1.

There are four main user network interfaces, R-, S-, T- and U- that provide the ISDN basic access at the physical layer. Out of these four interfaces, S and U interfaces are the most important ones. The S- interface is for transmission within customer premises and the U- interface for the connection of serving center to the customer premises. Figure 1.2 illustrates these two interfaces for a typical residential application.

The basic user-network U-interface standard recommendations given by CCITT suggests a 160kbps (2B + D) information rate for subscriber loops[2]. In the ISDN, the data rate at the U-interface is 144kbps. The U-interface is followed by the T-interface

BASIC ACCESS:

Rate: 192 Kbps

**Composition: B + B + D +
synchronisation and framing**

B: Information: Voice/Data

D: Signaling; or telemetry, packet

PRIMARY ACCESS:

B channels

Rate: 1.544/2.048 Mbps

Composition:

2.048 Mbps: 30B + 1D (each of 64Kbps)

1.544 Mbps: 23B + 1D (each of 64Kbps)

B: PCM Voice channel

D: Signaling

Figure 1.1: *The Channels for Basic and Primary Rate Access*

on the subscriber side, where the line rate is 192kbps. The extra 48kbps has been used at the T-interface for the purpose of housekeeping functions such as framing and balancing. Further, the T-interface is followed by the S-interface on the subscriber side at which the transmitting unit, the ISDN telephone is connected. While the S and T interfaces are well defined and have reached to final stage of standardisation, the U-interface still needs further study. We are presently interested in subscriber loops for digital transmission applications and we will call such a facility a *Digital Subscriber Loop*.

The *Standards Committee T1-Telecommunications, T1D1.3* [3] has proposed the line code for U-interface to be 2B1Q (2 binary 1 quaternary). This is a 4-level pulse amplitude modulation (PAM) code without redundancy. The bit-stream entering the NT (network termination) from the S/T interface (i.e., entering S/T interface towards the NT) is grouped into pairs of digits (bit fields) for conversion to quaternary symbols. Thus each successive pair of scrambled bits in the binary data stream is converted to a quaternary symbol to be output from the interface, as specified in Table 1.1.

The four values listed under "Quaternary Symbol" in the Table 1.1 should be understood as symbol names, not numerical values. The Figure 1.3 is an example of 2B1Q pulses over time. Square pulses are used only for convenience of display and do not in any way represent the specified shape of real 2B1Q pulses. Whereas the transmitted pulse shall have the shape specified in figure 1.4. The pulse mask for the four quaternary symbols shall be obtained by multiplying the normalised pulse mask shown in Figure 1.4 by 2.5 Volts, 0.833 Volt, —0.833 Volt or —2.5 Volt³. The nominal average power of a framed sequence of pulses with equal probability for each of the four symbols except during the synchronisation word is 13.5 dBm [3]. The symbol rate at the U-interface for the present usage will be considered as 80kbaud.

Now when we have looked at the basic technology involved with the U-interface, we also have to keep in our minds that the system should be efficient and cost-effective. To make a system cost-effective, several transmission techniques for the design of ISDN

³ It is recognised that for an interim period, until 1992, certain implementations may not be able to meet the 2.5 Volt pulse amplitude requirement. During this period between 2.5 Volts and 2.0 Volts will be acceptable[3].

Figure 1.2: T and U interfaces at user-premises

Table 1.1: Binary to Quaternary Conversion Table

First Bit (Sign)	Second Bit (Magnitude)	Quaternary Symbol (Quat)
1	0	+3
1	1	+1
0	1	-1
0	0	-3

U-interface have been explored by many investigators[4] to utilize existing two-wire twisted metallic pairs as a means of digital information carriers for full duplex data at 144kbps.

Various field trials and simulations carried out by operating telephone companies and other organisations show that out of the three important methods of transmitting data/voice over the digital carriers,

1. Frequency Division Multiplexing (FDM)
2. Time Division Multiplexing (TDM)
3. Echo Cancellation Technique (ECT) and Echo Cancellation Burst Mode(ECBM)

the *Echo-Cancellation (EC)* and *Echo-Cancellation Burst Mode (ECBM)* are found to be most promising [5][6]. In narrow bandwidth and power-limited environments, the EC technique is found to be of advantage. It is less affected by attenuation, transmission delay, crosstalk and impulse noise than the other methods. Though the EC technique is gaining more attention among most of the countries involved in the realisation of ISDN dream, Japan still supports time division multiplexing (TDM) and finds it to be useful for their local loop plant. We, in North America, will accept EC technique and this work also assumes the use of EC technique all over.

Figure 1.3: Example of 2B1Q Quaternary Pulses[3]

Figure 1.4: *Normalised Pulse From Network Termination appearing at Interface[3]*

Table 1.2: Table for Figure 1.4[3]

Places on the Pulse	Normalised Level	+3	+1	-1	-3
A	0.01	0.025V	0.00833V	-0.00833V	-0.025V
B	1.05	2.625V	0.8750V	-0.8750V	-2.625V
C	1.00	2.5V	0.833V	-0.833V	-2.5V
D	0.95	2.275V	0.79167V	-0.79167V	-2.275V
E	0.03	0.075V	0.025V	-0.025V	-0.075V
F	-0.01	-0.025V	-0.00833V	0.00833V	0.025V
G	-0.12	-0.3V	-0.1V	0.1V	0.3V
H	-0.05	0.125V	-0.042V	0.042V	0.125V

Crosstalk and impulse noise are likely to be the major impairments for higher (say, 144kbps) bit-rate transmission over two-wire loops, especially when the long reach of the digital subscriber loop is sought. It is the crosstalk coupling through which *impulse noise* gets coupled into digital information carriers. Some of the investigators [4] have already suggested impulse-noise models based on near-end crosstalk (NEXT) for echo-cancellation technique.

Figures 1.5(a) and 1.5(b) show the far-end crosstalk (FEXT) coupling path and the near-end crosstalk (NEXT) coupling path. In the synchronised burst-mode transmission (TCM), since the data is transmitted in one slot, the NEXT impairments will be less severe than the FEXT. Whereas, on the other side, with EC technique (also mentioned earlier), the FEXT will be less dominant than the NEXT because the far-end crosstalk power gets enough attenuated along the transmission line.

FEXT Coupling

$F^2(f)$: Far-end Coupling Power (a)

NEXT Coupling

$N^2(f)$: Near-end Coupling Power (b)

Figure 1.5: *Signal and FEXT model & Signal and NEXT model*

1.2 Problem Description

After going through the overview in the previous section, we get a basic idea of our telephone distribution plant. The presently used distribution plant was actually engineered for the low bandwidth of 3kHz, mainly to furnish the basic need of voice-band transmission. Now with the evolving integrated services digital network (ISDN), the basic access will be provided to our useful customers⁴ at higher bit-rate, 144kbps over ordinary copper wire pairs. In order to use the same loop-plant with higher data rate, it calls for highly sophisticated digital signal processors and also the most advanced & compact very-large scale integration (VLSI) circuits.

There might arise a question among the researchers, "*Why do we need to do so much research on copper-wire loop plant when the ultimate and better media for the transportation of high-speed digital data are fibers or coaxial-cables ?*" The answer to this is very simple; as we also mentioned earlier that most of the investment of transmission systems lies in local loop plants and therefore it will not be economically feasible to replace them before the end of their technical lifetime with fibers or coaxial-cables.

To use the same copper plant at considerably greater frequencies, the design of the digital subscriber loop (DSL) will vary considerably. Also we do not know many of the loop transmission characteristics at these intended frequencies. The phenomena at higher frequencies will be tremendously different from the present one. Among the major impairments like crosstalk, intersymbol-interference (ISI) and impulse noise, the models for crosstalk, based on frequency have been built [7]. ISI combatting has been made possible with the right choice of pulse-shaping filter and decision-feedback equalisation (DFE) but the characteristics of impulse-noise are still not known completely. Some investigators tried characterising impulse noise on 3kHz and 80kbaud data rate systems [8], [9], [4], [10] and some of these investigations do certify certain distributions of amplitude, time-of-arrival and duration of the impulse but still, before the first ISDN is shipped to the customers' premises, at least one or two years from now, some more

⁴ Businesses located in urban areas.

investigation need to be done to characterise impulse-noise. By looking at the facilities⁵ promised by the ISDN, we can easily foresee the class of basic users to be large business organisations and institutions. By the time, the ISDN services will be provided to residential users, it will be at least another quarter of the next decade. So, at present we need to concentrate on the systems which are going to serve businesses situated in dense urban environment and therefore for impulse noise, its characteristics should be studied in the same environment.

A recent measurement conducted by Bellcore⁶ in cooperation with NYNEX⁷ in New York and Boston, presented some time domain and frequency domain characteristics of impulse noise on the local loop plant [11]. The conclusion of this survey suggested that the impulse noise in the PBX environment should also be characterised. If high level impulses are found on rural and suburban areas, the performance will be severely degraded by the impulse noise. Therefore, since PBX loops could represent a more critical subpopulation, the work in this present study concentrates on a *PBX environment*.

We describe the *loop transmission characteristics* in brief here as they affect high speed digital transmission. These transmission characteristics are found to be varying with the frequency and the transmitted pulse's magnitude gets attenuated (depending upon various features of the loop plant) and also the pulse-width gets elongated as it travels on the transmission lines. Particularly at high bit-rate transmission, the closely spaced symbols suffer the phenomenon of *intersymbol interference (ISI)*; this can lead to errors at the receive-end.

If the loop make-ups are known accurately, these errors on the digital subscriber loop can be predicted with more accuracy. The terminology, loop make-up stands for the construction of the particular lines as far as the length of sections, mixture of different wire-gauges and the bridged-taps are concerned. The loop make-up keeps changing with the demand and the need of the location that particular loop is serving. Therefore, the

⁵ There is a mention of this in section 1.1 of this work.

⁶ Bell Communications Research, U. S. A.

⁷ A telephone operating company in New York and New England.

operating telephone companies will have to maintain better records of changing loop plants which in-turn can help designing the system by predicting the errors.

In this thesis we are concerned with the impulse noise characterisation in particular, for an impulse noise event:

- (i) What is its shape?
- (ii) How big does it become?
- (iii) How long does it survive?
- (iv) How often does it come?
- (v) Where and when does it appear on loops?
- (vi) Where does most of the energy lie in frequency domain?
- (vii) Could a certain source of origin be predicted?

These will be described in the following sub-sections.

1.2.1 What is its shape?

The impulse noise events were captured by a measurement system designed for this purpose only. A preset threshold was set to look for the events at least as big as that threshold. A system was designed to capture and record impulse noise events in such a way that whenever the impulse noise crosses the predefined threshold level, it (the system) starts to digitise the line-voltage pulse with some predefined delay (time axis off-set). Each event recorded, returns 1024 samples of digitised data. The minimum threshold level was decided by considering the peak transmitted voltage from the CCITT recommendations [3] and the worst-case line-loss for the worst-loop⁸ (we are considering loop # 9 for this work, see Figure 1.6) and then the received pulse shape was drawn for the 50% duty cycle, 6.25 μ s (half the symbol period⁹) wide transmitted pulse of peak voltage 2.5 Volts. Figure 1.7 shows the 50% duty cycle square pulse. Thus the

⁸ The worst loop is defined as the loop#3 of loop-plant which has maximum number of irregularities like bridged-taps, bridged-taps of bridged-taps, different wire-gauges and also the length of the loop [3].

⁹ For 80kbaud transmission of 2B1Q data, the symbol period will be 12.5 μ s, half the symbol period will then be 6.25 μ s.

minimum threshold level will be the received pulse height. The measurement system design and methodology are defined in details in chapter 3.

After going through the captured impulse-noise records, we find that most of the pulses have different pulse shapes, some of them resemble simple sinusoid and some of them are like a burst of pulses (Figure 1.8 and Figure 1.9). Some of the events are without any regularly known shapes, like a dying-out pulse (Figure 1.10). These pulses will be discussed in more detail in the later chapters dealing with the time-domain and frequency domain analysis of the impulsive events.

1.2.2 How big does it become?

During the measurements, it was found that the magnitude of the impulses varies a lot with the time of day and also some times, the pulses seemed to depend upon some other activities in the exchange. These activities could be because of central-office (exchange) switching transients or may be, because of a high frequency modem carrier on a neighbouring wire pair. The pulse height was also dependent on the threshold set-up for the measurement. Different thresholds were set up to observe the magnitude of the pulses. The magnitude of the pulses varies from less than 5mV to greater than 35mV. But most of the pulses were in the range of 3mV—20mV and a very small proportion of the pulses were found to be at or above 35mV. This will once again be dealt with in detail in the later chapters.

1.2.3 How long does it Survive?

The impulsive events' bank (a file storing all the windowed¹⁰ events) shows that most of the impulses survived for 40 — 70 μ s. Some of the impulses had a single lobe, some had double lobes as if two pulses were superimposing on each other and there were multi-lobe pulses too which seemed to occurred from a ringing kind of disturbance on the line.

For a transmission speed of 80kbaud, the symbol period at the receiving-end, after the equalisation will be 12.5 μ s which is far less than some of the durations of the

¹⁰ The windowing of the received pulses is described in Chapter 4.

Figure 1.6: Loop # 3 and Loop # 9 from the loop-plans.

Figure 1.7: *50 % Duty cycle pulse.*

Figure 1.8: PULSE — 1

Figure 1.9: PULSE — 2

Figure 1.10: PULSE — 3

observed pulses. This shows that for high speed transmission over copper wires, the impulse noise could cause multiple errors in the received symbol.

1.2.4 How often does it Come?

The arrival rate of impulse noise events was measured with respect to the threshold levels. For the lower threshold value, the occurrence rate of impulsive noise was higher in comparison to the occurrence at higher threshold levels. The threshold level was not the only criterion, the time of day was another factor on which arrival rate depended. In most of the measurements, the cumulative distribution of the time-of-occurrence of the pulses with respect to the absolute time of day is found to be very consistent over the ensemble of the events on different pairs and at different days. This shows the human habit of using telephones in a business environment.

The average time between the pulses did not seem to vary considerably with the change in pairs whereas it did depend upon the threshold levels. The number of impulses observed in 30 minutes varied between 150 and 1200.

1.2.5 When and Where does it appear?

When: The measurements carried out by us on subscriber lines connected to a PBX (at Mitel Corporation, Kanata, Canada) show that the arrival of the impulse noise was quite uniform throughout the weekday. We want to emphasize here that this study of impulse noise is mainly in the PBX loop plant environment, in contrast to other studies of central office environments. Most of the measurements were carried out during the weekdays and within the office-hours, i.e. from 9:00 AM to 5:00 PM. In our measurements, a particular loop was monitored at a particular threshold for the whole day. This was helpful in averaging the arrival time statistics at the certain set threshold. This way all the thresholds were tried at each pair for the period of at least 7–8 hours.

Where: During the period of measurements, the type of cables were changed from all data carrying cables to all voice carrying cables and also to the ones which carry voice/data together. The occurrence of impulsive noise was studied at these cables for the initial period and it was noticed that there was not much of difference in the statistics

of amplitude, time of arrival and duration of impulses. Eventually, it was decided to that we take measurements only on the cables which carry voice/data together.

1.2.6 Where does most of the energy lie in Frequency domain?

The records of the impulses were processed¹¹ to see the frequency-domain characteristics of the events. Within the frequency of interest, it was seen that most of the energy was situated in the lower end of frequency spectrum i.e., upto 200—300kHz. Above that the energy was seen to be highly attenuated. Some of the frequency plots had peaks which were found to be caused by some regular activity on the adjacent pair. This tone was removed from the recording and then the captured event's frequency characteristics were seen. Still, most of the events reflected their most of energy being concentrated at lower frequencies.

Only a small percentage of events had their energy situated at higher end. This could once again be true because the current induced in the pair under measurement is due to the high-pass or low-pass action of the cable-plant or also because the sources producing the events(for example, relay closures) are at relatively low frequency. Most of the cable-plants act as high-pass filter. In Figures 1.11 and 1.12 two examples of frequency domain characteristics are shown for the sake of reference. Such curves will be shown and analysed in the frequency domain analysis in later chapters.

1.2.7 Could a certain source of Origin be predicted?

Can we find a certain source of Impulse Noise occurrence on the digital subscriber lines, other than switching and bursty activities in the exchange? Can a certain transmission scheme used in voice and data band modems, functioning on the neighbouring cables, be another cause of such bursty disturbance? If we can predict such kind of information from the measurements, it may be easier for a communication system designer to construct a scheme to combat the impulse noise.

¹¹ The captured raw data was passed through 4th-order Butterworth Filter at two cut-off frequencies. The 3dB cut-off frequency for 80kbaud system was chosen to be 40kHz and for the 400kbaud system it was 350kHz.

Figure 1.11: *Frequency domain sample of a pulse.*

Figure 1.12: *Another example of Frequency domain Characteristics*

Combating of Impulse Noise can also be attempted from a different angle, which is; if one can find a “typical-event” in time-domain from the ensemble of the data and also a “typical-frequency curve” from the same ensemble of data, and then do the reverse, such that, find “typical-frequency curve” from the “typical-event” and a “typical-event” from the “typical-frequency curve”. Now if both the “typicals” match upto some extent, we can call that “typical-event” as “characteristic impulse” and then design a filter which will be capable of combating this “characteristic impulse”! Some investigators in the Netherlands have tried such a study [12] but they did not use enough data for finding a typical-pulse. They have suggested the use of well known smearing and desmearing filters to protect the system from this kind of noise.

Once we have answered these questions we will know the various characteristics of the impulse noise on the digital subscriber line.

1.3 Thesis Organisation

The thesis is organised into six chapters.

In Chapter 2, we review the characteristics of subscriber loops, as they affect the performance of digital transmission over of copper wire pairs.

Chapter 3 illustrates the design aspects of impulse noise measurement system. It also explains how the measuring equipment is controlled by software via an IEEE-488 bus.

Chapter 4, the core of this work, presents the analysis of the measured and re-constructed data. This chapter also displays few examples of impulsive events. These examples mainly include raw voltage-surges, filtered events at different cut-off frequencies and their frequency domain characteristics.

In Chapter 5, the author has attempted to evaluate the performance of a digital subscriber loop system in the presence of such a bursty nature of noise and the ever-present Gaussian Noise.

Finally, in Chapter 6, we have discussed impulse noise on the 2-wire pairs and also suggested extensions for the future work in this area.

Chapter 2

Channel Constraints

2.1 Noise on the Digital Subscriber Loop

Whether the existing loop plant is capable of transporting digital signals depends upon various factors which impair the performance of the digital subscriber loop (DSL). Mainly, the impairments enumerated below are the prime factors in deciding the performance of a loop plant [13].

1. Attenuation
2. Inductive Loading
3. Different Gauges
4. Bridged Taps
5. Noise

Noise is one of the major performance deciding factors. Here, in this work, we only talk about different type of noises with an emphasis on Crosstalk. The most important types of noise which degrade the performance of the DSL are listed below:

1. Crosstalk
2. Impulse Noise

3. Internal System Noise

We throw some light on the above listed types of noise and discuss their effect on digital subscriber loops. Crosstalk is being considered here because impulse noise is found to be coupled via crosstalk paths [14].

2.1.1 Crosstalk

This is one of the impairments to digital transmission over copper loops. In the cables consisting of large numbers of metallic pair lines, crosstalk among the adjacent lines at high bit-rates transmission, may be a dominating origin of disturbance. As already mentioned in the previous chapter crosstalk is classified into two types, namely:

1. Near-End Crosstalk (NEXT)
2. Far-End Crosstalk (FEXT)

The maximum transmitted power of a digital system is given by the compatibility requirements with other systems like program channel services, Digital Data Phone (DDS) or Dataroute, as well as analog carrier (SLC—8) and digital carrier such as T1 [15]. In order to ensure the performance of other systems being interfered with by such a noise, the transmit level should be limited within some range. This, in the case of an analog carrier system, is done by ensuring that the idle noise does not exceed the transmit level. Whereas, in the digital carrier systems, this is done by ensuring that the eye-pattern does not close sufficiently at the decision point to cause the bit-error rate (BER) to exceed the requirements. The minimum possible received signal power is governed by the neighbouring systems' crosstalk into the digital subscriber lines. An attempt is made in digital carrier transmission to keep the crosstalk level low enough not to cause BER requirements to degrade. This is possible only when the characteristics of crosstalk are known on the DSL.

The decibel (dB) difference between the peak-transmitted signal level and the minimum received-signal level helps in deciding a maximum-bound of range for the DSL system. The NEXT and the FEXT also contribute to the limitations of the system range. The outcome of this is the maximum signal loss that the DSL system can

tolerate without the level becoming so low with respect to crosstalk that bit-error rate requirements are violated. The minimum of these two ranges represents the actual system range without any bounds due to impulse noise. The statistics of impulse noise found in the previous studies for high bit-rate transmission [11], [12], [16] over 2-wire loops, shows that in the ISDN environment, impulse-noise could cause severe problems, especially for data transmission.

The NEXT and FEXT models and corresponding mathematical formulae involved in calculating the pair-to-pair crosstalk coupling loss are given as follows.

2.1.1.1 Far-End Crosstalk (FEXT)

Far-end crosstalk is the coupling between the cable conductors of the signals transmitted in the same direction. This can be seen in Figure 2.1. The far-end coupling path, between any two conductors, in the same bandwidth, is on the average given by the equation (2.1):

$$|F(f)|^2 = K_0 \left(\frac{l}{l_0} \right) \left(\frac{f}{f_0} \right)^2 \quad (2.1)$$

where K_0 is a random variable defined as the gain between two cable conductors measured at the same initial length l_0 and the initial frequency f_0 . Equation (2.1) gives the received power of a transmitted-signal of frequency f divided into the coupled power of the same signal received on a neighbouring conductor pair of the identical length, l . Expression (2.1) is multiplied by the transmitted signal's power spectrum of the interfering FEXT entering the cable. Since, the given ratio is based on the received power, equation (2.1) can be rewritten in equation (2.2) in decibels units and is called the *Equal Level Coupling Loss*:

$$ELCL = \left[10 \log K_0 + 10 \log \left(\frac{l}{l_0} \right) + 20 \log \left(\frac{f}{f_0} \right) \right] \quad (2.2)$$

Equation (2.2) shows that the FEXT coupling loss decreases at the rate 6dB/octave with frequency and at the rate 3dB/octave with conductor length l . Figure 2.1(a) shows

two conductors in the same cable, with signals directed in the same direction as the other, having response $C(f)$.

By making use of the Figure 2(a), we can write the power of the single system coupled into another system. The average FEXT noise power P_f at the output of the receiver-2 is given as equation (2.3):

$$P_f = \int_0^W |G(f)|^2 |C(f)|^2 |R(f)|^2 |F(f)|^2 df \quad (2.3)$$

where,

$G|f|^2$ = PSD of the disturbing system

$C(f)$ = Cable response

$R(f)$ = Receiver gain of the disturbed system

W = System bandwidth

Pair-to-pair FEXT interference noise in dB:

$$P_F = m_F + 10 \log \left(\frac{l}{l_0} \right) + 10 \log \int_0^W |G(f)|^2 |C(f)|^2 |R(f)|^2 \left(\frac{f}{f_0} \right)^2 df \quad (2.4)$$

where m_F is mean pair-to-pair far-end coupling loss for a cable of length L [7].

2.1.1.2 Near-End Crosstalk (NEXT)

The interference of strong near-end transmitted signals on weak received signals give rise to near-end crosstalk (NEXT). The NEXT is the coupling between cable conductors of the signals transmitted in the opposite direction. The near-end coupling loss (NCL) between two transmission paths inside the cable bindings is expressed on the average by the power ratio given by the equation (2.5).

$$|N(f)|^2 = K_0 \left(\frac{f}{f_0} \right)^{\left(\frac{3}{2}\right)} \quad (2.5)$$

In decibels (dB) this can be presented as:

$$NCL = 10 \log K_0 + 15 \log \left(\frac{f}{f_0} \right) \quad (2.6)$$

FEXT Coupling

$F^2(f)$: Far-end Coupling Power (a)

NEXT Coupling

$N^2(f)$: Near-end Coupling Power (b)

Figure 2.1: *FEXT and NEXT models and their coupling.*

The value $(\log K_0)$ is a random variable which expresses the coupling loss between the conductors under consideration measured at the initial frequency f_0 . From the equation (2.5), we find that NEXT coupling loss decreases at the rate of 4.5dB/octave with frequency. Figure 1.1 also shows the coupling path between the two transmission lines transmitting in the opposite direction to each other. The average NEXT noise power at the output of the receiver-2 can be given by the equation (2.7):

$$P_N = \int_0^W |G(f)|^2 |R(f)|^2 N^2(f) df \quad (2.7)$$

In the same way we can write pair-to-pair NEXT interfering noise in dB:

$$P_N = 10 \log \int_0^W \left(\frac{f}{f_0} \right)^{\left(\frac{1}{2} \right)} |G(f)|^2 |R(f)|^2 df \quad (2.8)$$

On the basis of some investigations, only NEXT need be considered when determining the capability of digital systems and other systems using the echo-cancellation technique (ECT) and sharing the same facility. NEXT interference from T1 into the digital system will also have to be considered seriously [15].

2.1.2 Impulse Noise

Impulse Noise consists of short duration pulses of unwanted signals which occur in random pattern in the transmission medium. In the telephone networks, within the range of frequency of interest, transients lasting for a few milliseconds or microseconds and generated by switching operations constitute the largest source of impulse noise. The major sources of the impulse noise are the dial pulsing transients, switching and high voltage ringing-pulses in the central office (CO).

Impulsive noise which is predicted to be originated by the causes given above, mainly the relay operation, is found to be coupled to the signals on the digital carriers, in the same cable quad or the adjacent cable¹², using NEXT paths [4]. It is understood

¹² Effective only when the disturbance is very strong.

that with the advent of all-digital facilities in all over North-America¹³, the effect of impulse noise will eventually be negligible. But this will not be the case for the coming couple of decades. And for that period, it becomes necessary to ensure that digital transmission in the loop-plants does not get corrupted by impulsive noise.

Presently, from whatever information we have regarding the characteristics of impulse noise on loops for voice-band, it is very hard to come up with a certain model of impulsive noise. But, by doing the direct measurements on the loops situated at different geographical locations, we can understand the phenomenon of impulse noise in a better way and then it will be easier to come up with some useful statistics of impulse noise.

Recently, some investigators have done measurements of impulse noise for high-bit rate [10], some of them did the measurements in the laboratories on simulated cables and/or real-cables [4] and they suggested different models for the time-of-arrival, amplitude of the impulses and for the duration of the pulses. Except for one investigator [10], no one suggested the amplitude distribution to be *Gaussian*. Pierre Mertz [17] suggested the hyperbolic distribution for amplitude of the impulse noise assuming the arrival to be *Poisson*; the GTE investigations [10] suggested the amplitude distribution to be *Gaussian* or *two-sided exponential* and arrival to be *Poisson*. The study conducted at Nippon Telephone and Telegraph (NTT), Japan [18] claimed the amplitude distribution to be *Logarithmic Normal Distribution*. The inside-laboratory measurements conducted at Australian Telecom Research Laboratories, Australia [4] give a theoretical background for the error-rates at high-bit rate transmission in an impulsive noise environment.

Earlier, we said that the impulses get coupled to the neighbouring digital-data carrying lines through near-end crosstalk paths (NEXT), it has also been found that the relay (the main cause of origin of impulse noise) comes in series with the NEXT path [4]. The near-end coupling acts as a filter whose impulse response can be given as $c(t)$ and the corresponding transfer function as $C(f)$. This transfer function can be

¹³ Our present domain of interest.

given as [7]:

$$C(f) = \sqrt{K_N f^{\frac{1}{2}}} \quad (2.9)$$

where, the coupling constant $K_N = 0.6 \times 10^{-13}$.

The block diagram of Figure 2.2 shows the cascading of the relay operation and the NEXT path filter with its transfer function, $C(f)$.

The input to the NEXT filter will be in the form of a step-function and this step function over the ensemble of adjacent cables gets differentiated to give pulse-like output which in turn acts as impulses. The differentiation phenomena in the frequency domain can be represented by $R(f)$ and be given by:

$$R(f) = \sum_{i=0}^{\infty} a_i e^{-j2\pi f \tau_i} \quad (2.10)$$

where, a_i 's are the amplitudes of the arriving impulses and τ_i 's are the impulse arrival instances, see Figure 2.3.

Therefore, the response of the cascading of the relay and the NEXT filter can be denoted by $N(f)$ and is given as:

$$N(f) = R(f) C(f) \quad (2.11)$$

the equation (2.11) is the basis of understanding impulse noise mitigation in the digital subscriber lines (DSL).

More characteristics of impulse noise will be presented in Chapter 4 and its effect on a DSL transmission system in Chapter 5.

Relay and NEXT Filter Cascading

Figure 2.2: Relay and NEXT Filter Cascading for Impulse Noise Modelling

Impulse Arrival Distribution in Time Domain

$t_i \Rightarrow$ Arrival Instances
 $dt \Rightarrow$ Intra-impulse arrival time
 (Poisson Distributed)

Figure 2.3: *Impulse Arrival Distribution in Time Domain.*

Chapter 3

Measurement System Design and Methodology

3.1 Hardware

This section will explain some of the important factors which were considered in choosing the hardware for the design of the measurement system.

The measurement system used in this work was made by using a combination of commercially available equipments and specifically designed interface units to monitor the induced voltages appeared on a pair (one under measurement) in a multi-pair cable. The analog voltages were converted into digitised samples and these samples were stored on the hard-disk of the controlling computer for later data-screening and analysis.

The complete equipment setup for the measurement of impulsive noise on the loops is shown in Figure 3.1. A cable bundle was chosen at random from the ensemble of cables used to support a PABX¹⁴ facility. This cable-bundle carried eight smaller cable packs (let us call the smaller cable-packs as quads, also see Figure 3.2) inside. Each

¹⁴ Private Automatic Branch Exchanges.

of these quads contained 25 copper-pairs. The cable plant is divided into three types of services, namely:

- Only voice carrying cables.
- Only data carrying cables.
- Voice/data carrying cables.

A pair was selected from inside of the inner-most quad in first two type of the cables. Let us denote the pair under measurement as *disturbed loop*.

A network termination (NT) of 100Ω was used on both the ends of the disturbed loop. The selection of 100Ω termination instead of 600Ω (for the existing loops) was made because the nominal driving point impedance at the interface looking towards the network termination (NT) for high bit-rate system is found to be about 135Ω [3]. The resistance in these measurements was varied between 90 to 120Ω and hardly any difference was noticed, finally we decided upon 100Ω for our purposes. The measurements carried out at NYNEX by Bellcore used 120Ω at both ends of the loop. No other active sources of voltage were on the disturbed cable while measurements were being carried out.

The analog voltage measured at the end of the loop-pair across the 100Ω resistance was digitised simultaneously and stored in the RAM (random access memory) of the 80286 processor based controlling-computer. The controller was programmed in such a way that whenever the voltage appearing on the disturbed line crossed a certain preset threshold, a snap-shot of the digitiser input was taken. In order to store an impulse along with preceding time axis information, a short delay (time axis off-set) was specified during the measurement. The terminology, 'snap-shot' means; at the moment of the impulse hit whatever voltage waveform is displayed on the digitiser screen is divided into 1024 quantised samples and these quantised samples are stored in the RAM of the controller. The quantised samples were stored as 16-bit (words) with sign bit to be always positive (0). The negative voltages were also represented by the positive quantised samples. All these quantised samples (of positive and negative impulses)

Figure 3.1: Measurement Equipment Set-Up.

Each Quad carries 25 twisted copper pairs.

Figure 3.2: *Cable Bundle and Quad*

were later re-constructed using the calibration formula¹⁵ to give actual voltage and time axes values of the waveforms stored. Using this information the captured data was stored for further analysis.

3.1.1 Calibration of the Digitiser

For the purpose of digitising the analog voltage waveforms appearing on the disturbed loop, we made use of a commercially available, Hewlett-Packard programmable digitising oscilloscope, model hp54200a/d. This oscilloscope had a facility to be controlled by a computer via an IEEE-488 bus. Another feature of great interest to our measurements, was voltage threshold. This feature allowed us to set a threshold level and do the pulse counting as well as recording. The digitiser had bandwidth of 50MHz, which was enough for our measurements.

Calibration of the digitiser was another important task for the analysis of digitally stored waveforms. The digitiser is set up to send the waveform data after a read request is received from the controlling software. Upon request, it (the digitiser) sends waveforms in words of 2 bytes. The total waveform has 1001 points, each represented by a word, thus resulting in a total transfer of 2002 bytes every read. The selected Y-axis range is quantised into 128 levels for a waveform whereas the digitiser sends values ranging from 0 to 123 (124 levels), called *quantisation levels*. There was found to be a bug while calibrating the digitiser, consequently we have used 124 quantised levels for our analog-to-digital conversion. These values are converted to voltages using information obtained from the scope through the *preamble?*¹⁶ sub-command of the oscilloscope. The following program segment (a part of 'C'-code) shows a usage of *ieeewt*, *ieeerd*¹⁷ and *preamble* commands.

Program Segment 1:

```
ieeewt("output01; run\n");
ieeewt("output01; wav\n");
```

¹⁵ Please see Program Segment-2 in section 3.1.1 for this formula.

¹⁶ Hewlett-Packard's 'hp54200a/d' commands.

¹⁷ The IEEE-488 bus controlling routines. ©IOTech, Ohio.

```

        ieeewt("output01; form word");
        ieeewt(";sour 1\n");
        .
        .
        ieeewt("output01; pre?\n");
        ieeewt("time out 0\n");
        ieeewt("enter01\n");
        ieeerd( preambled );
sscanf(preambled,"%d,%*d,%d,%*d,%lg,%lg,%*d,%lg,
%lg,%lg,%*d,%*s",&numb,&xin,&yin,&yoff,&yref);
        .
        .
        ieeewt("output01; data?\n");
        ieeewt("enter01\n");
        ieeerd( curve );

```

In the previous program segment 1, a pre? is sent to the digitiser and the controller keeps waiting for the digitisers' reply. The digitiser's reply to the controller's request is then stored in a variable called preambled (see line 8 of the program segment 1). The preamble command returns useful information; this information is enumerated in the following example.

```

Format
Type
Points
Count
Xincrement
Xorigin
Xreference
Yincrement
Yorigin
Yreference
Coupling
Label

```

Most of these values were supplied by the user during the course of digitiser initialisation. By taking these values from the digitiser's preamble query makes sure that

the supplied values are correctly being taken by the digitiser. The formula used for the conversion of raw sample values to actual voltage values is shown in program segment 2.

Program Segment 2:

```

halfrange = yin * yref;
factor     = halfrange/62; /* (2^7-4) */
correction= yoff - halfrange;
.
.
amp_axis_vector = raw_sample_vector * factor + correction;

```

3.1.2 Threshold of the Detection Equipment

In most of the earlier investigations for impulse noise measurements and characteristics, the impulsive voltage was passed through various bandwidth filters. [11], [12]. It is true that in an actual digital subscriber loop (DSL) system, the receive filter comes before the decision is made for the received symbol, and this fact justifies the use of filter between the point of detection and the storage medium. For the present measurements, our philosophy is little wider in nature. That is, once the received voltage has been band limited by filtering, there is no way that we can get the original impulsive voltages from the filtered stored-data, whereas if we have stored the unfiltered (or wideband filtered, say a bandwidth of 2MHz) data, we can always let this stored-data pass through well designed software-implemented digital or equivalent analog filters of any kind and order. For our analysis, we have passed the data through a fourth order Butterworth filter of 50kHz and 350kHz cutoff for ease of analysis at two bit-rates. But if there is a need, one can always change the type, the order and/or the bandwidth of the filter realised by the software and do the analysis.

The threshold of the detection equipment is another parameter to be set while measuring the impulsive disturbances on the disturbed loop. In order to find a minimum threshold above which the data symbols may be in error, we can use the various channel characteristics based upon the CCITT recommendations for high bit-rate transmission

over the two-wire copper loop. Following channel characteristics are used to decide the minimum threshold level of the detector:

- ☐ Transmission rate of symbols — 80kb/s (160kbps).
- ☐ A pulse having 100% duty-cycle, 2.5 Volt peak voltage is transmitted from the transmitter end.
- ☐ Channel used is the Bellcore loop # 9 [3].
- ☐ The received pulse is drawn in the time domain to see the line attenuation.

So, if we consider a line attenuation of 45dB (say for the worst case) at the frequency of interest, we can calculate received pulse amplitude magnitude with respect to the transmitted pulse amplitude of 2.5Volts. Figure 3.3. If $V_{in} = 2.5$ Volts and $g_{dB} = -45$ dB then the V_{out} can be calculated to be 14mV (without the receive filter). If the filter is taken into account the filter attenuation will change the receive pulse magnitude.

So, the minimum magnitude considered was 5mV and then we chose a few other threshold levels to classify the recorded data. These thresholds are as listed below:

- ☐ 5mV
- ☐ 10mV
- ☐ 20mV
- ☐ 35mV

The measuring equipment was then tested in a laboratory environment by generating the sharp impulses using an impulse generator as well as telephone cradle switches (the pulses were fed through a current-probe in this case). The bandwidth of measurement and the detection levels were also varied to check the proper functioning of the controlling software and the IEEE bus.

A complete strategy behind the measurements will be described in the next section of this chapter.

Figure 3.3: *Output of the Loop#9 with respect to the 100% duty cycle input.*

3.2 Software to Control the Equipment

The programmable digitiser used in our measurements was controlled using a 80286/12MHz based central processing unit (CPU). The various features of the digitiser could be brought into use by supplying the different commands from the CPU. An IEEE-488 interface was used to connect the CPU of the computer and the programmable analog-to-digital converter.

A program in Microsoft-C^{TM18} (version 5.1) was generated on a personal computer and was run to pass the various parameters to control the digitiser functions. There were three versions of computer programs made, one of them was on-line query type, the next one was command-line argument type and the third one was made to observe the time of occurrence statistics alone. The on-line query version is useful for novice users and for more skilled users, the command-line program is provided to pass the parameters to the digitiser through an IEEE-488 bus.

It was made sure while generating the 'C'-code that the program could be abandoned anytime during the execution. This was possible by just hitting any alpha-numeric key on the keyboard.

The program also displays all the measurement parameters (complete setup) just before the execution and if they are not correct, the program could be stopped by using ^C (Press C—key while control-key is pressed) interrupt of MSDOS^{TM19} system. The following program segment gives a partial C routine to stop the program during execution:

Program Segment 3:

```
do
{
  c = _bios_keybrd(_KEYBRD_READY);
  if(c != 0)
  {
    c = _bios_keybrd(_KEYBRD_READ);
```

¹⁸ Microsoft-CTM is a Trade—Mark of Microsoft Inc.

¹⁹ MSDOSTM is a Trade Mark of Microsoft Inc.

```

        printf("c : %s\n", c);
        if(c != '\n') goto closeprg;
    };
    ieeewt("output01;rdy?\n");
    ieeewt("enter01\n");
    ieeerd(rdybuf);
    sscanf(rdybuf,"%d", &status);
    printf("Status = %x\n", status);
} while ((status & 4) == 0);
.
.
closeprg: _SAVE_EXIT;
.
.

```

The measurement parameters could also be supplied by editing a file named oscilpar.dat in the current directory where the executable measurement program is residing. An example of such file showing important parameters for the measurements can be shown as in program segment 4.

Program Segment 4:

Loop Number	: V1-RGR-123
Triggerlevel	: 10e-3.
Volt range	: 40e-3.
Volt offset	: 0.0
Time range	: 1e-3.
Time offset	: 100e-6.

The program also requires name of a file where the captured impulse noise data is to be saved. The name of the storage file is supplied either on the command line without any extension. The file extensions are suffixed by the program later, depending on the count of pulses. Thus, for example, if the storage file name is 'feb23da1', then at the end of execution of the program, files like feb23da1.1, feb23da1.2, feb23da1.3 etc. will be found in the current directory. The .1, .2, .3 and so on extensions will stand for first, second, third and so on impulsive voltages captured. This way one can supply different names of data storage files for different sessions of

measurements.

After each session of measurement, the data was backed-up on the high-density floppy-diskettes and space was made available for next measurements. Each 1001 samples of impulsive noise waveform was stored on the hard-disk in the form of an ASCII text (in a file). Each file containing information about the time and voltage axes as well as the time of occurrence of an impulse, consumes approximately 28kbytes of hard-disk space. The hard disk of total space of 40Mbytes, was partitioned into two drives namely C: and D:. The C: drive was used to store all the system software, source code, executable files, editors and temporary data files. The D: drive was strictly used for data storage and temporary backups. Finally, the data-format was converted to UNIX^{TM20} operating system by passing it through dos2unix²¹ filter and transferred to SUN-network through the network file server (NFSTM)²² facility at our Radio Communications Laboratory for further reconstruction and graphical analysis of the impulses on the high-density graphical Sun workstations.

A simplified pseudo-code of the software for measurement can be shown here in the following program segment 5 (pseudo-code). In this pseudo-code presented here, the main flow of the algorithm is illustrated in bold letters and at few places a partial 'C'-code is also presented for the ease of understanding the algorithm.

Pseudo-Code:

/* Pseudo-Code for the Controlling Software */

Input duration of pulse count storage on command line.

Open and Read 'oscilpar.dat' for measurement parameters into computer memory.

²⁰ UNIXTM is a Trade Mark of AT&T.

²¹ A NFSTM utility.

²² NFSTM is Trade Mark of SUN Microsystems Inc., the makers of SUN family of computers.

```

if((oscilparameter=fopen("oscilpar.dat","r"))==NULL)
{
    printf("Unable to open %s for reading!!\n",
           oscilparameter);
    exit(0);
}
rewind(oscilparameter);

```

Scan the contents of oscilpar.dat in the variables and close it.

Initialise the IEEE-488 bus for the measurement session

```

if(ieeeinit() == -1)
{
    printf("Cannot Initialise IEEE488 bus card !!\n");
}
ieeewt("clear01\n");

```

Reset the Digitiser (the hp54200a/d) and establish connection with the remote host.

```

ieeewt("output01; rst\n");
ieeewt("remote01\n");

```

Check for any keyboard hit.

```

if(anykeybrd hit)
{
    goto close program after saving whatever has
    been received
}
else
{
    continue further;
}

```

Supply the parameters from the memory to the digitiser via IEEE-488 bus.

Wait for the digitiser to send a signal when a pulse above a threshold appears on the line.

```

if(NO)
{
    go back to look for an impulsive waveform;
}
elseif(YES)
{
    continue further;
}

```

If signalled YES from the previous check, store the captured impulse noise data in the RAM.

Decrement the pulse count

Check if the number of pulse are zero yet?

```

if(NO)
{
    go back to the step where digitiser
    initialisation is done;
}
elseif(YES)
{
    continue further;
}

```

Open a file in the same directory with the supplied filename with extension.

```

for(j=1; j <= desired_pulse_count; j++)
{
    sprintf(wavedata,"%s.%d",input_file,j);

if((stream = fopen(wavedata,"w")) == NULL)
{
    printf("Unable to open %s for writing!!\n",
        wavedata);
    exit(0);
}

```

Write all the information about the impulse. The important information which will be needed later during the course of analysis are:

File Name:
Impulse #:
Loop # :
Triggerlevel:
Volt range :
Volt offset :
Time range :
Time offset :

Close the above file.

Decrement the pulse count.

Check if pulse count is zero.

```
if(pulse_count != 0)
{
    go back to the beginning of FOR LOOP;
}
elseif(NO)
{
    continue further;
}
```

Display number of pulse counted.

End

Chapter 4

Analysis of Measured Data

Before the measurements of impulse-noise in a PBX loop plant were started, a primary run of measurements were carried out. In this primary run, we chose four different cable bundles, supporting different services on the lines. These services were Voice Communication, Data Transmission and Voice/Data Transmission. One cable at a time was considered and all the four were monitored for 360 hours (approximately 24 hours and more for each) distributed over a period of 60 days at different thresholds and constant bandwidth of measurement (same sampling rate). The bandwidth of the measuring equipment was chosen just wide enough to utilize the measurements to cover services up to 800kbps. This bandwidth was 2MSamples/second. Our main aim of the primary run was to gain an idea about the nature of impulses occurring on each of the four cables. The result of this showed us that there is not much difference in the impulse noise occurrence on those cables. Therefore, we decided to stick to an idle cable whose surrounding cables supported data/voice services together.

Following this primary run of measurements, we started our main measurements on the chosen pair which was also surrounded by very busy loop-pairs.

The data analysis sections involve examination of noise waveforms, power spectra, relative power spectra, time of occurrence, duration of impulses, amplitude and time of

occurrence distributions.

4.1 Examination of Waveforms

This stage mainly deals with visual examination and classification of the time-domain information of the recorded waveforms. During the course of measurement 200 waveforms were collected and stored in different files for easy access later for the analysis. The waveform pictures were classified by their shapes. Within each class no two waveforms were found to be exactly the same in shape and duration; this shows that there is no fixed recurring activity on the adjacent pairs causing a voltage spike to occur, except for the bursty-pulses (too many adjacent spikes in a recording), which seemed to look similar, were kept in a separate class. That was because sinusoidal-looking pulses were hard to differentiate from each other. An example of such waveform is shown in Figure 4.1. These bursty-pulses, after getting filtered through the cable (acting as a high-pass filter most of the time) become one type of impulsive noise occurring on the disturbed lines. These pulses lie in the middle of the spectra and usually are high enough to cause symbol in errors for longer loops.

4.2 Windowing the Data

Some of the waveforms dominated by a spike-like pulse, plus the system background noise, were passed through a box-car rectangular window to minimize the effect of the background noise on the computed spectra. While windowing the waveforms for better understanding of the disturbing events, we faced a problem of window size. It was not wise to keep a constant size of window. By keeping a constant size, we might have

included some of the background noise in the impulsive event, or
truncated the long duration events even before they die out.

To rectify both of the above problems of sizing the window, a software was generated in Pro-MATLAB (a signal analysis language on Sun-OS4), which uses adaptive window size depending upon the shape of the impulsive event. The adaptive window adjusts

Figure 4.1: *An example of Bursty-pulse recording on the adjacent lines.*

its width depending upon where the peak of the impulses occurs. During the course of windowing, it was observed that the width of the window varied from 60 to 200 samples. The window duration can be calculated by using the following formula.

$$\text{Event Width} = \text{Window width} \times \text{Sampling Period} \quad (4.1)$$

If the sampling period and the width of the rectangular window (in number of samples count) are known, the duration of the impulsive event could be found. These windowed waveforms, here onwards, will be called *impulsive events* or simply *events*.

These events are stored separately in a file or different files each containing one event. The advantage of doing this is that we could use these events (non-filtered) and pass them through some low-pass filter of interest and use the filtered events for frequency domain analysis as well as in DSL system performance evaluation (Chapter 5 deals with DSL system performance).

4.3 Spectra of Waveforms

The energy spectrum of each unwindowed waveform was displayed to see in which frequency range most of the power resides. Since, this (unwindowed) data contained background noise and quantisation noise (because of digitiser), after taking FFT of this waveform, the frequency domain curve was found to be full of spikes. The energy spectrum of the unwindowed data was calculated as follows.

If x is a vector containing the calibrated²³ samples corresponding to the amplitude of an impulse and n is the length of this vector, then the FFT of this sample vector could be written as:

$$X(k+1) = \sum_{n=0}^{N-1} x(n+1) W_N^{kn} \quad (4.2)$$

where,

²³ Obtained by using the correction factor formula given in Section 3.1.1, Program Segment:2.

Impulse Event Width = Window width x Sampling Period
 if Sampling Period = $(500 \times 10^{-6})/1024$

$$= 0.5 \times 10^{-6} \text{ seconds/sample}$$

and Window width = 60, then

Impulse Event width = 30 microseconds

Figure 4.2: *Windowing of a Waveform.*

$$W_N = e^{-j(\frac{2\pi}{N})}$$

N = length of the vector 'x'

The energy spectrum, a measurement of the energy contained in the signal at various frequencies is then calculated by multiplying the fourier transformed output with its conjugate. Therefore, if X is the FFT of x , then the energy spectrum denoted by P_{xx} will be given by,

$$P_{xx} = X.X^* \quad (4.4)$$

where X^* is the conjugate of X . For the sake of simplicity and understanding, a program segment generated using MATLAB is presented here to calculate PSD.

Program Segment:1

```
time_axis_vector = 0:time_axis_range/1024:time_axis_range;
amp_axis         = eval([data_file_var(i,:), '(8:1031)']);
amp_axis_vector  = amp_axis*factor + correction;

no_of_samples = 1024;

Y = fft(amp_axis_vector);
% 'fft' is function call made to calculate fast fourier transform
% of the amplitude vector.
%
% The Energy_spectrum, a measurement of the energy
% at various frequencies, is:

energy_spectrum = Y.* conj(Y);

% 'conj' finds the conjugate of the complex quantity
% energy_spectrum.

energy_spectrum512 = energy_spectrum(1:512);
```

```

% Now form a frequency axis for the first 512 points (The other 512
% points will be symmetrical.)
%
    sampling_freq = (1/time_axis_range)*no_of_samples;
%
% Here we are taking only 512 points out of total
% 'no_of_samples' (1024). Suppose a sequence of N points
% (1024 here) is obtained at a sample frequency of
% fs(sampling_freq here). Then, for up to the Nyquist
% freq., or point n = N/2+1, the relationship between the
% bin_number (on the frequency axis) and the actual frequency is:
%
%       Actual_freq = (bin_number -1)*fs/N
%
%(ref: 3-44 of Matlab Manual)

    freq_axis_vector = (0:511)*sampling_freq/no_of_samples;

% Now we plot the energy_spectrum512 Vs. Frequency!

    plot(freq_axis_vector, energy_spectrum512);

```

Figures 4.3 and 4.4 are presented here as examples of the frequency domain analysis of the unwindowed impulse noise data.

As we mentioned earlier, the frequency domain curves were full of spikes (see, for example Figure 4.3), it was therefore hard to obtain anything useful out of those curves. Therefore we decided to display relative energy spectrum with larger frequency bins. Examples of relative energy spectra are shown in Figures 4.5 and 4.6. A simple algorithm was used to find relative energy spectrum. The algorithm is as shown below:

- (a.) Add total energy in the given bandwidth.
- (b.) Make 20 bands of total bandwidth each of 50kHz.
- (c.) Add all the powers in each band.
- (d.) Divide (c.) by (a.), which returns relative power of that band.

Figure 4.3: *Energy Spectrum Curve for an impulse recording (unwindowed).*

Figure 4.4: *Energy Spectrum for another impulse recording (unwindowed).*

The above algorithm could be given mathematically by,

$$\begin{aligned}
 P_T &= \sum_{i=1}^N P_N \\
 P_b &= \sum_{j=m+1}^{m+\frac{N}{n}} P_j \\
 P_{rel} &= \frac{P_b}{P_T}
 \end{aligned} \tag{4.5}$$

where,

$$\begin{aligned}
 m = \text{step}(k) &\implies k \text{th value of array 'step'} \\
 k &= 1, 2, \dots, n \\
 \text{step} &= 0, \frac{N}{n}, \frac{2N}{n}, \dots, nN
 \end{aligned}$$

The P_T , P_b and P_{rel} are the total power in the given bandwidth, the individual band power and the relative energy spectrum, respectively.

After going through the relative energy spectra of different waveforms, we realised that the background noise causes many little bumps in the curve which really distracts attention during the course of analysis. To avoid this, we used the windowed waveforms in the frequency domain analysis. Figures 4.10 and 4.11 show the energy spectra of the windowed events.

Now, the FFT of each event was taken and energy spectrum was displayed. These curves are found to be very informative in a sense that whatever frequency spectra we were seeing was due solely to impulsive events, not due to the background noise. Like the previous investigations of impulsive noise at different locations and different exchanges, these frequency spectra seemed to be concentrated in the lower frequency region upto 250kHz and the power spectra were pretty flat in this frequency region.

For the sake of better understanding, a few more frequency spectra of the unwindowed events are displayed in Figures 4.7 through 4.9. If we observe the various spectra

- (a.) Add total energy in the given bandwidth.
- (b.) Make 20 bands of total bandwidth each of 50kHz.
- (c.) Add all the powers in each band.
- (d.) Divide (c) by (a), which returns relative power of that band.

Figure 4.5: *A Relative Energy Spectrum Calculation of unwindowed data*

Figure 4.6: *Another Relative Energy Spectrum example of unwindowed data*

Figure 4.7: *Energy Spectrum of another Impulse recording (unwindowed data).*

Figure 4.8: *Energy Spectrum of Another impulse(unwindowed data).*

Figure 4.9: *Energy Spectrum of Another impulse (unwindowed data).*

Figure 4.10: *Energy Spectrum of a windowed waveform.*

Figure 4.11: *Energy Spectrum of another windowed waveform.*

presented here, we find that most of the impulsive activity falls in the lower portion of the spectra, that is from 0 to 250kHz, whereas the NEXT spectra falls in the upper portion (NEXT is caused due to the high-pass phenomena of cable).

4.4 Filtering the Events

As mentioned in the earlier chapter on measurements, we did not use a filter during the measurements. This was done to enable us to store raw-voltage waveforms to apply in the analysis of different bit-rate systems. But in a real-life data transmission and reception systems, we do make use of receive filter (usually a low-pass filter to block the noise-band). Therefore, while analysing the data for 160kbps and 800kbps systems, we pass each of the impulsive events through a 4th order, low-pass Butterworth filter. This filter has been used separately for both the bit-rates. The filter 3dB cut-off frequency for a 160kbps (80kbaud) system was chosen to be 40kHz and for the 800kbps (400kbaud), it was 350kHz.

After passing each event through the filter, the smoothed output events were again studied in the frequency-domain. Again the FFT of these filtered events was taken and it was found, that even after passing the events through 40kHz cutoff filter, most of the energy was situated in 0-200kHz band. Examples of this are displayed in Figures 4.12 and 4.13

4.5 Impulsive Events vs. Time of Day

From the collected data at different thresholds, histograms have been drawn to illustrate the percentage of impulse noise events versus time of day. Since most of the measurements were carried out during office hours on weekdays, the time of day ranged from 9:30 AM to 5:00 PM. Figures 4.14, 4.15, 4.16, 4.17 are the four histograms drawn from four different measurements and at different thresholds (specified on each of the graphs). As we increase the threshold levels, the number of impulsive-hits generally becomes smaller. As a matter of fact, this phenomenon was found to be true during the entire course of measurement, but at one point (see Figure 4.15, of 10mV threshold)

we noticed that during the period when most of the other threshold levels were calm (on other days), the 10mV level had lots of impulsive-hits. At 5mV level, one day, between 9:30 AM to 10:00 AM, there were 1200 impulsive hits whereas on another day during the same time of the day at higher threshold, 10mV there were only 151 impulsive occurrences. Such statistics can be generalised if a larger amount of data is collected over a long period of time and then analysed.

4.6 Inter-arrival Times

Figures 4.18, 4.19, 4.20, 4.21 show the histograms of the percentage inter-arrivals versus inter-arrival time (seconds). These are once again drawn with respect to different thresholds (specified on each figure). After going through all these histograms (dotted lines), it seems most of the inter-arrivals range from 1.3 to 1.5 seconds. Because of the limitations imposed by the clock rate on the personal computer and the IEEE bus, the minimum measurable intra-arrival time was 1.3 seconds. To avoid this problem from affecting the various distribution (discussed in the later sections), the windowing of data approach was applied. Still, for more accurate inter-arrival time statistics hardware would have to be developed which make use of assembler and interfaces with GPIB²⁴ or HPIB²⁵ and also have provisions for measurement bandwidth and timer-settings. Other than this drawback, the hardware design by using commercially available discrete equipments was well-suited for the such measurements. More measurements need to be carried out which will be dedicated to only inter-arrival time statistics.

Table 4.1 gives a quick glance of mean inter-arrival times and the corresponding threshold levels. We can see clearly that the arrival time changed substantially by changing the threshold level from 20mV to 35mV. The increase in inter-arrival time by increasing the threshold level is obvious because of the fact that the digitiser was triggered less frequently than it would have for lower threshold levels. Once again, by having a large number of measurements, it would be easier to generalise these statistics.

²⁴ General Purpose Interface Bus

²⁵ Hewlett Packard Interface Bus

Figure 4.12: A Frequency domain plot of filtered event.

Figure 4.13: Another Frequency domain plot of filtered event.

Figure 4.14: *% of Impulse Events Vs. Time of Day (at 5mV)*

Figure 4.15: % of Impulse Events Vs. Time of Day (at 10mV)

Figure 4.16: *% of Impulse Events Vs. Time of Day (at 20mV)*

Figure 4.17: % of Impulse Events Vs. Time of Day (at 35mV)

Table 4.1: Threshold Levels and Mean Arrival Time

Threshold levels	Mean Arrival Time
5mV	1.354 Sec
10mV	2.235 Sec
20mV	1.745 Sec
35mV	38.75 Sec

But, even if we have small amount of measurement data, it (the data) matches with the behaviour of impulse arrivals at different thresholds.

In the study carried out by Bellcore for NYNEX in the Boston and Manhattan area they have indicated the mean arrival time to be in seconds at central office. For instance, at 2mV threshold level, 500kHz bandwidth, in New York Central Office location, the mean intra-arrival time is presented to be 13.27 seconds. Similarly, on other threshold levels of detection, the mean inter-arrival time has been found to be generally in the order of seconds. If we compare this with our results obtained and being presented here, we see that at 5mV threshold level, the mean arrival time is 1.354 seconds. The answer to this may be; in a PBX environment where the length of the loop under measurement is of the order of few hundreds of feet (approximately 300 feet), the signal strength will be high enough to avoid any errors caused due to noise. Therefore, the design criteria for the loop hardware may be less stringent, which in turn could mean that these noise levels (35mV) are not high enough to cause any errors.

4.7 Threshold versus Impulsive Events

It would be a good idea to see the effect of the threshold level on the occurrence of impulsive events. For this, we have drawn a histogram of the percentage of impulsive events versus threshold levels in Figure 4.22. It is interesting to notice in Figure 4.22 that first it shows the dependency of the number of pulse-hits on the threshold levels and secondly, it also shows that even after changing the threshold from 10mV to 20mV,

Figure 4.18: % inter-arrivals Vs. Inter-arrival Time (at 5mV)

Figure 4.19: % Inter-arrivals Vs. Inter-arrival Time (at 10mV)

Figure 4.20: % Inter-arrivals Vs. Inter-arrival Time (at 20mV)

Figure 4.21: % Inter-arrivals Vs. Inter-arrival Time (at 35mV)

Table 4.2: Count of Impulses, Thresholds and Percentage of Impulses

Count of Impulses	Threshold Levels	Percentage of Impulses
17511	5 - 10mV	33.94%
15516	10 - 20mV	30.08%
15091	20 - 35mV	29.25%
3469	$\geq 35\text{mV}$	6.72%

there was not much of difference in impulsive occurrences. However, by changing the threshold from 20mV to 35mV, the percentage of impulsive hits was very much lower. This gives us an indication that the pulses arriving on the loops were between 10mV and 20mV most of time. As said earlier, the impulse arrival decreased (intra-arrival time increased) by increasing the threshold level because the oscilloscope was not triggered that often.

The Table 4.2 shows the relation between the Count of Impulses, Threshold Levels and Percentage of Impulses. Each threshold level was monitored for a whole day from 9:30 AM to 5:00 PM

4.8 Amplitude and Duration of the Impulses

In this section we look into the characteristics of the impulses from the *duration and amplitude* side. By probing the statistics of the amplitude and duration of the impulses, we will be able to apply these results to the system performance analysis (to be done in the next chapter).

Figure 4.22: *Cumulative % of Impulse Events Vs. Threshold (mV)*

4.8.1 Impulse Noise Amplitude Distribution

In order to get a feel of variation in impulse noise amplitudes with respect to various threshold levels, we present a statistical analysis leading us to the conditional amplitude distribution. Before we start to describe the conditional amplitude distribution of impulsive events, we will introduce the correct interpretation of the term *Conditional Amplitude Distribution*.

To calculate the conditional amplitude distribution, we find the fraction of time the impulsive amplitude is greater than a threshold level given the condition that an impulse hits (occurs) at that threshold level. To put these words into an easily recognised mathematical formula, we show the following:

$$\text{Conditional Amplitude Distribution} = \frac{\sum_{i=1}^N \tau_i}{T} \quad (4.7)$$

where,

$i = 1, \dots, N$

τ_i = duration of each pulse at the given threshold

T = Total duration of observation (window width)

N = Number of Pulses above threshold

Each impulse noise recording before any filtering was picked up and the duration of all the impulses at the different threshold levels were summed-up separately for the total duration of observation. By doing this for the different threshold levels; 2, 5, 8, 10, 12, 15, 18, 20, 22, 25, 28, 30, 32 and 35mV(6.02, 13.98, 18.06,, 30.1, 30.88 dBmV), we find the behaviour of variation of pulses with respect to threshold levels. For an example of setting different threshold levels on a recording (of impulses), Figure 4.23 is displayed which is a randomly chosen impulse and a threshold level at 5mV is drawn on this. Similarly, other threshold levels are also drawn later and the conditional amplitude distribution statistics is calculated. The τ_i 's (as mentioned earlier

in Equation (4.7)) and T are marked on the figure for the better understanding of the methodology. Once again, the T is the total time of each recording (observation of each recording). This duration is in milli-seconds range.

The computed conditional amplitude probabilities at all the threshold levels were plotted against the corresponding threshold levels. The Figure 4.24 plots the probabilities of amplitudes being greater than abscissae given that a pulse occurs for all the recordings individually (marked by '+' on the plots). A mean of all the probabilities is also drawn to get an average variation of an ordinate with respect to the abscissa. This is shown by a bold line with '*' (stars). It was felt at this point that mean distribution could be biased because of too large or too small pulse amplitudes among the ensemble. Therefore, the median conditional amplitude distribution was also plotted to cancel the effect of too large or too small amplitudes. This is also shown by a bold line with 'o'.

The sole purpose of this amplitude analysis is to enable a communication system designer to predict the effect of impulsive noise on the digital subscriber loop system for high bit-rate voice/data transmission. As a small example; if the receiver threshold is set at 18dBmV, it is found from the mean amplitude distribution that 12% of the bits will be in error given that an impulse strikes.

4.8.2 Duration Distribution of Impulsive Events

In assessing the effect of impulse noise on high bit-rate transmission systems, the distribution of the duration of pulses is another important factor. The impulsive noise event duration will determine the duration of error events due to impulsive noise.

The duration of impulsive events is the time for which the impulsive noise waveform remains above the threshold level of interest. This is calculated by drawing a line at the threshold level of interest and calculating the time difference between the beginning and the end of the impulse. Figure 4.25 presents it more clearly.

The windowed events (windowing as described in section 4.2) were filtered through a 4th order low-pass Butterworth filter at cut-off frequencies of 40kHz and 350kHz. The filtered events at these cut-off frequencies were tested separately for 80kbaud and

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Figure 4.24: *Amplitude Distribution Vs. Threshold Levels*

Figure 4.24: *Amplitude Distribution Vs. Threshold Levels*

400kbaud bit-rate system. The duration of the impulses was calculated at 14 threshold levels (2, 5, 8,....., 32, 35 mV) for each event at both the bit-rates. The 40kHz filtered events correspond to a 80kbaud system whereas 350kHz ones correspond to a 400kbaud system.

The calculated distribution of impulse noise events results at 80 and 400kbauds bit-rate were sketched with respect to the threshold levels on abscissa. For individual events at different thresholds, a scatter plot was made (marked with '+' s). Mean and median of duration were also drawn at the corresponding threshold levels (marked with bold lines). These results are displayed in Figure 4.26 and Figure 4.27 for 160kbps and 800kbps bit-rate systems, respectively.

The duration of impulsive noise events analysis will help communication system designers in predicting the number of symbols affected whenever there is an impulsive noise transient. For more understanding of this analysis results, we present a simple example for a 80kbaud bit-rate system. For the 80kbaud transmission rate using the 2B1Q code (4 level PAM) each symbol period will be $1/80 \times 10^{-3} = 12.5\mu\text{s}$ and for the 400kbaud transmission rate the symbol period will be $1/400 \times 10^{-3} = 2.5\mu\text{s}$. Therefore if an impulse hits during the transmission and its' mean duration above receiver threshold (detector) level is $20\mu\text{s}$ (say), which will mean that impulse event cover a duration of $1\frac{1}{2}$ symbol period for 80kbaud system and 8 symbol period for 400kbaud system. This in turn, also suggests that impulsive noise will prove to be more trouble some for higher bit-rate systems than 80kbaud ones.

Now by observing mean and median duration curves in the Figures 4.26 and 4.27, we calculate the average and median impulse noise duration, given the receiver detector level (threshold) and an impulse hit occurs. If we assume the threshold level to be 18dBmV, the average duration of events is found to be $18\mu\text{s}$ for 160kbps transmission rate and $30\mu\text{s}$ for 800kbps system. By applying the above example here, we approximately calculate that for 160kbps system the average duration will cover 1.5 symbol periods, whereas for 800kbps system, the same statistics will be 12 symbol periods.

Figure 4.26: *Duration of Events Vs. Threshold, 160kbps*

Figure 4.27: Duration of Events Vs. Threshold, 800kbps

For the case of in-depth understanding at higher threshold level ($> 18\text{dBmV}$), the duration of events versus threshold level plots have also been represented with logarithmic ordinates for 160kbps and 800kbps in Figures 4.28 and 4.29.

4.8.3 Number of Crossings

The impulsive amplitude distribution and the distribution of duration above corresponding threshold levels, statistics are less informative and incomplete without the number of crossings of impulsive events at those threshold levels statistics. The amplitude distribution information will be more useful, if one knows the amplitude distribution of an impulse noise recording at a threshold as well as knowing how many times the pulses cross that threshold level. This is true because of the fact that we would know how many impulses are above that threshold. Similarly, in the case of the duration of impulsive events information at a threshold level, it will be of more use if it is known how many times the events cross that threshold level. Once again the reasoning is same as above.

The number of crossings of impulse noise during the total time of observation (the number of positively crossing impulses) are counted. This procedure is repeated 15 times (including no threshold for zero length loop) for all the threshold (0, 2, 5, 8, 32, 35 mV) and the statistical data is stored in arrays (or files) for later use with previous two statistics. Figure 4.30 explains this graphically.

All the impulsive noise recordings were tested for the calculation of numbers of crossings at 15 different thresholds (as mentioned earlier) and the resulting data is plotted in Figure 4.31 ordinate with respect to all the threshold levels on abscissa. The results of individual recordings are plotted as a scatter plot (marked '+') and mean & median are sketched with bold lines with 'o' and '*' legends respectively. The logarithmic ordinate curve has also been presented here.

If we know the conditional amplitude distribution at a certain receiver detector threshold level, and we also know how many times the impulses cross that level during

Figure 4.28: *Duration of Events Vs. Threshold (Logarithmic Ordinate), 160kbps*

Figure 4.29: *Duration of Events Vs. Threshold Levels (Logarithmic ordinate), 800kbps*

T = Total time of observation

Figure 4.30: *Definition of Number of Crossings w. r. t Threshold levels*

Figure 4.31: *Linear Number of Crossings Vs. Threshold levels*

Figure 4.32: *Logarithmic Number of Crossings Vs. Thrshold levels*

the total time of observation, we can estimate how many impulsive lobes have arrived during that time. These lobes could be because of only one excitation of impulsive disturbance or because of more than one.

4.9 Typical Pulse Shapes

By looking at the ensemble of the windowed events from different measurements, we find out that the shapes of the impulses are varying from time to time. But at these impulsive voltages, he can see that the classification of the impulses can be made depending upon the shape of the pulses as described below.

While going through the ensemble of the events, we saw that the set of events could be divided into three major groups. The classifications into groups will solely depend upon the shape of the pulses in each event. The three groups are defined as follows:

- ☐ Positive going single cycle type events.
- ☐ Negative going single cycle type events.
- ☐ More than one prominent cycles-events type (positive and negative both considered).

Keeping the above definitions in mind, the total set of events was divided into three parts. Let us call these groups as A^+ , A^- and AA^+ , respectively. Figures 4.33, 4.34, 4.35 show the events belonging to the Group- A^+ , Figures 4.36, 4.37, 4.38 show the events belonging to the Group- A^- and the Figures 4.39, 4.40 illustrate the events of Group- AA^+ .

The main aim of grouping the different events in three groups was to see if a few characteristic pulse shapes consistently recur. If they do, system designers might think of a measure to combat impulsive noise on the digital subscriber loops. One could come up with a question, *"Why don't we find one characteristic pulse which will represent the whole ensemble of the events?"* It is true that if we have one characteristic pulse representing the set of events, we can devise a comparatively simple method to combat the impulse noise on DSL. Use of smearing and desmearing filters has been found to

be an effective method to combat the impulse noise [12](but only when we have one characteristic pulse for all). But by looking at the ensemble of events, it was hard to decide upon a characteristic pulse, since, all positive, negative and positive-negative pulses were mixed.

Seeing all these difficulties in deciding the typical pulse shape, we grouped the events and then came up with three typical pulses by averaging the each group as formed earlier. The Figure 4.41 shows the typical pulse generated from the Group-A⁺. If we notice this typical pulse, we find that it has resulted in two-pulses in succession. This has happened because of the orientation of the pulses. If the grouping is done more precisely, such that the multi-lobe pulses are placed in the Group-AA⁺ and more events are used to find the average, the resultant typical pulse will be single—lobe type and dominating in the positive direction.

The Figure 4.42 shows the typical pulse corresponding to the Group-A⁻. This typical pulse also has two pulses in it, but the second pulse is very low in amplitude and can be ignored.

The Figure 4.43 typical pulse represents the Group-AA⁺ and the shape is of the multi-lobe type.

Figure 4.33: *Group-A+ Events*

Figure 4.34: *Group-A+ Events (continue from previous figure)*

Figure 4.35: *Group-A+ Events (continue from previous figure)*

Figure 4.36: Group-A- Events

Figure 4.37: Group A- Events (continue from previous figure)

Figure 4.38: *Group A- Events (continue from previous)*

Figure 4.39: Group AA+ Events

Figure 4.40: Group AA+ Events (continue from previous)

Figure 4.41: *Typical Pulse of Group A+*

Figure 4.42: *Typical Pulse Group A-*

Figure 4.43: *Typical Pulse from Group AA+*

Chapter 5

DSL System Performance in Impulse Noise Environment

With the increasing demand for high bit-rate voice and data transmission, a need has arisen to evaluate the digital system performance in the presence of constraints like impulse-noise, crosstalk, echo, jitter and bridged taps.

The U-interface bit-rate in ISDN has evolved from 80kbits/sec to 160kbits/sec and 1.544Mbits/sec. The 160kbits/sec transmission must be reliable for two-wire loops upto 3.5 Miles(=5.833Km), in the presence of major impairments as mentioned above. However, in order to cater the growing demands for the communication facilities, modems with bit-rates up to or above 1.544Mbits/sec and longer loops (> 3.5 miles) would be needed.

For the sake of completeness, the measured and analysed impulse-noise data will be used in this chapter with a simulated digital subscriber loop system to investigate its performance at high bit-rates, 80kbaud and 400kbaud. Instead of using simulated impulse-noise for the BER and SNR performance of a DSL system using a linear and an adaptive decision feedback equaliser in the receiver structure, the reproduced digitised impulse-noise events and gaussian noise are being used here as a noise generator.

To derive maximum achievable bit-rates at which one can transmit successfully, channel capacity or its upper bounds can be derived. Therefore channel capacity could be considered as a performance parameter. By using more complex modulation and timing recovery schemes, close to the theoretical channel capacity could be achieved.

The major parameters affecting the performance of a digital subscriber loop system in a band-limited and power-restricted environment are found to be ISI, crosstalk, impulse-noise and additive white gaussian noise (AWGN). In this chapter, we will evaluate the distribution of the number of errors per impulse noise event on a DSL system in the presence of gaussian and impulsive noise, assuming there is no crosstalk. The AWG noise is generated by simulation with variance (σ^2) = 1×10^{-13} , whereas the impulsive noise generator block has been created by making use of real-time measured data. These measurements, as mentioned in the previous chapters, were made in a PBX loop-plant of MITEL Corporation, Kanata, Canada. A simple 2B1Q (4 level PAM) transmitter and adaptive DFE receiver have been utilised in the simulated DSL system.

5.1 System Architecture

The system architecture considered here for the performance evaluation uses basic digital systems blocks like binary signal generator, an encoder, channel, AGC, receive filters, equalisers, sampler and a detector. These blocks are shown in Figure 5.1. The Echo-Canceller is not shown in this figure for the sake of simplicity. The hybrids are included in the channel response, one at each side of the channel block.

The input sequence $\{a_i\}$ is a binary sequence representing data to be transmitted. This sequence is operated upon by the 2B1Q (4-Level PAM) encoder to produce output sequence $\{c_i\}$. No pulse-shaping is employed on this low-pass signal. The encoded data bits are then passed through a channel²⁶ having impulse response as $h(t)$ and the corresponding frequency response $H(f)$. Consequently, the received signal at the front-end of the receiver can be represented as:

$$x(t) = \sum_{i=0}^{\infty} c_i h(t - iT) + (n_{imp}(t) + n_{gauss}(t)) \quad (5.1)$$

In the above equation, $n_{imp}(t)$ represents the impulsive noise whereas $n_{gauss}(t)$ represents the gaussian noise.

²⁶ Channels considered here are from the BELLCORE loop plant. See Appendix-A for their configuration.

The received signal is first passed through a filter, a linear equaliser and then sampled at a symbol rate $1/T$ samples per second. The sampled received signal, y_n can be given by,

$$y_n = \sum_{j=0}^L a_{n-j} h_j + \nu_n \quad (5.2)$$

where, a_n 's are the transmitted data symbols. The h_j 's are the samples of the overall impulse response. The impulsive and the gaussian noise samples are represented by ν_n 's. They are uncorrelated with data symbols.

In the following sub-sections, the important system blocks and their importance in an impulse noise affected digital subscriber copper loop system will be discussed.

5.1.1 The 2B1Q Encoder

The binary data to 2B1Q mapping is explained in brief in the Chapter 1 of this work. We will throw a light on the choice of 2B1Q code for the North American Integrated Services Digital Networks. Many of the investigators have reported the superiority of 2B1Q scheme over binary and other multi-level codes. The reference [19] discusses the advantage and compares the performance of a DSL system using 2B1Q codes. over other codes.

The main issues in selecting the 2B1Q code were the Impulse Noise and the NEXT impairments [19]. The NEXT and the FEXT²⁷ of a transmission system are dependent upon the number of states (levels) in the code, its power spectral density (PSD), the modulation, transmit & receive filters, the kind of equalisation and the detection method. The 2B1Q at 80kbaud suffers less attenuation than binary transmission at 160kbaud with the difference increasing with loop length. Typically, the two codes differ by about 0.5dB/kft of 26AWG cable. Therefore, the zero redundant quaternary code, the 2B1Q code, is identified by the ANSI Standards Committee to be a good compromise that maximizes both performance and implementation simplicity.

²⁷ Not severe in the Echo-Canceller techniques.

5.1.2 DSL Channel Response Simulation

Simulation of the transmission channel makes use of impulse responses in the time domain for the Bellcore loop-plant listed in the Appendices-A. This Bellcore loop-plant was devised and later adopted by the T1D1 committee to test the performance of the received signal. A computer program was developed on Sun-OS4 for the calculation of channel response. The simulation was based on transmission lines and the two-port networks theory.

Subscriber loops in the telephone loop-plant may consist of sections with different gauges and may also contain bridge taps or bridge taps of bridged taps. A model using hybrid or ABCD parameters can be used for transmission lines. This is shown as in Figure 5.2(a) and the same two-port network can be written in the form of a matrix as follows:

$$\begin{bmatrix} v_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} v_2 \\ I_2 \end{bmatrix} \quad (5.3)$$

where,

$$A = \cosh \gamma l$$

$$B = Z_0 \sinh \gamma l$$

$$C = \sinh \gamma l$$

$$D = A$$

$$\gamma = \text{propagation constant} = \sqrt{(R + j2\pi fL)(G + j2\pi fC)}$$

$$Z_0 = \text{characteristic impedance} = \sqrt{(R + j2\pi fL)/(G + j2\pi fC)}$$

$$R = \text{Resistance}$$

$$L = \text{Inductance}$$

$$G = C$$

In order to calculate the complex A, B, C and D parameters, accurate values of γ versus frequency, f and Z_0 versus frequency for various gauges, temperatures and cable types had been taken from the appendix of reference [20].

Digital Subscriber Loop system block used in the simulation

Figure 5.1: The Digital Subscriber Loop System Architecture

(a)

(b)

(c)

Figure 5.2: ABCD Parameters (a) Two port network, (b) Application of ABCD parameters, (c) A line with a open ended bridged tap.

For a given line with no bridge tap, see Figure 5.2(b), a voltage gain at a particular frequency is given by:

$$H = \frac{1}{Z_s \left(C + \frac{D}{Z_L} \right) + A + \frac{B}{Z_L}} \quad (5.4)$$

Let us consider another loop with a single bridge tap, see Figure 5.2(c). If A', B', C', D' are the ABCD parameters for the open circuited bridge tap, the equivalent ABCD parameters of the bridge tap can be stated as :

$$\begin{bmatrix} A = 1 & B = 0 \\ C = Y_o = \tanh \gamma_b & D = 1 \end{bmatrix} \quad (5.5)$$

where, $Y_o = \text{admittance} = \sqrt{\frac{C'D'}{A'B'}} = \frac{1}{Z_o}$ and γ_b is propagation constant for the entire bridged tap, i.e., the complex log of half the sum of $A', B'/Z_o, Z_o C', D'$.

In this manner one can find ABCD parameters even for more complex lines having different configurations. Some of the impulse responses generated using the ABCD theory have been shown in Figures 5.3 and 5.4 for two different lines.

The channel in time domain has been simulated as a transversal filter with 240 taps which is equivalent to $240/8 = 30$ symbol intervals for 8 samples/symbol. As mentioned earlier, the Bellcore loops are used for these simulations. The lengths, gauges and bridged taps vary from Loop#1 through Loop#15 and during the simulation these will be tested to see whether they pass the performance test (described later) in the presence of only Gaussian and Gaussian + Impulsive noise or not. The configuration of these loops has been tabulated in Table 5.1, and Appendix-A shows their physical configuration.

The *performance test* of the loops consists of counting errors in the detected samples. The errors caused during the training mode will not be counted towards the transmission errors.

Simple hybrids, each at transmitting and receiving ends of the 2-wire loop have been used. Figure 5.5 shows the construction of this hybrid. It has been a common practice of different U-interface designers to use different kind of circuitry for the hybrids to suit

their requirements for better system performance. The one used here is a simple hybrid of impedance-balancing type of structure.

The noise added to the signal is considered to be White Gaussian with zero mean and desired variance, σ^2 plus the impulse noise. To generate exponentially distributed impulses, a computer program was developed in Unix-'C' which generates exponentially distributed time periods and includes measured impulsive events randomly at those intervals. The impulse noise events are added on top of the embedded background gaussian noise. The program is made to give enough flexibility to change the number of impulses to be included, Lambda (λ , the mean arrival time) to be changed, and also let the user supply an output file name for the generated data storage. For the sake of convenience, the user is suggested to use the filename, `imp_noise.mat`. The extension, `.mat` is suggested here for the ease in loading the generated file on to the Pro-MATLAB²⁸ workspace and then observe it graphically on the SUN's high resolution screen.

The distribution of the gaussian noise samples were generated using Unix system's `rand( )` function which generates uniformly distributed random numbers. These uniformly distributed numbers were then converted to the gaussian distribution by using a `sin( )`, `cos( )` algorithm. A program segment shown below makes the algorithm more clear to understand. Most of the samples were found to be uncorrelated.

Program Segment :

```
VARIANCE = 1e-13;
MEAN = 0.0;
count = 0;
jk = 0;
ex_counter = 0.0;
for (kk=1; kk <= interval[array_len]; ++k)
{
    uul = (((double)rand())/large_num));
```

²⁸ The MATLAB signal processing package was used to see the maximum and minimum amplitudes of the impulses and the embedded gaussian noise, graphically.

Figure 5.3: *Loop response for the Loop#1 from Bellcore loop-plant*

Figure 5.4: *Loop response for the Loop#2 from Belcore loop-palm*

Figure 5.5: *An impedance-balancing hybrid structure*

```

uu2 = (((double)rand())/large_num));
uullog = log10(uu1);
if (jk == 0)
{
    ty[kk]=(sqrt(-2.0*uullog))*(cos(2.0*pi*uu2))*gain_factor;
    jk = 1;
}
else
{
    ty[kk]=(sqrt(-2.0*uullog))*(sin(2.0*pi*uu2))*gain_factor;
    jk = 0;
}
count += 1;
ty[kk] = ty[kk] * sqrt(VARIANCE) + MEAN;
tylpf[kk] = lowpass_filter(ty[kk])*gauss_gain;
    /* Low-Pass Filter */
}
(void)printf("Count of background_noise array = %d\n", count);

```

A receiver filter is a 4th order lowpass Butterworth filter used with two cutoff frequencies, 40kHz and 350kHz for the two bit-rates, 160kbits/sec and 800kbits/sec. The purpose of this filter is to eliminate out of band noise. The Butterworth filter coefficients were generated by using Pro-MATLAB on Sun network and these coefficients were imported by the main simulation in Unix-'C'.

5.1.3 Equalisation

After filtering the out of band noise, the receiver processes the signal with the equaliser. The equaliser plays an important role in the DSL system. An adaptive equaliser becomes necessary to restore the transmitted pulse-shape at the front-end of the equalisers. The $\sqrt{f}$ attenuation causes pulse dispersion over the length of the two-wire copper loop. This problem becomes more severe with the increasing loop lengths, varying wire gauges and the inclusion of bridged-taps. The use of linear equalisers is one choice to undo the pulse broadening at the receiving end. Though these linear equalisers are widely used in the commercially available voiceband modems, certain drawbacks make their implementation more difficult on a VLSI circuit. To name the

major drawbacks, complex multiplications and noise enhancement are two of them. However adaptive equalisation without using the multipliers can be achieved with a decision feedback equaliser (DFE). The noise enhancement problem is also reduced due to the inherent nonlinear property of the DFE which, in turn means, the great reduction in complexity and implementation. A major drawback of DFE is that it is only capable of removing inter-symbol interference (ISI) of the trailing type that is, within its span. Therefore all other disturbances including precursor and ISI have to be suppressed well in advance before they start to propagate errors in the DFE and make it impossible to converge. To achieve this a $(1-\mu D)$ filter (fixed linear-equaliser or differentiator) is used before the decision feedback equaliser. This equaliser is kept active over a broad range of cable lengths.

The adaptive LMS algorithm, used for the DFE coefficients $C_n(i)$ is:

$$C_n(i+1) = C_n(i) + \gamma \times \text{error} \times a_{i-m} \quad (5.6)$$

where,

$m = 1, \dots, N$

N = Number of taps

a_i = Transmitted data bits.

Gain, $g(i)$ is also adapted using the same algorithm as above:

$$g(i+1) = g(i) + \gamma \times \text{error} \times a_i \quad (5.7)$$

where, γ is a step-function used for fine adjustments. For our simulations this quantity was kept to be small enough not to cause any convergence errors in the DFE. The Figures 5.6 shows the linear fixed equaliser (a differentiator).

5.2 Simulation to Assess ISDN System Performance

In the preceding chapters we have explained how the measurement equipment was designed and how the measurements were carried out. After measurements, the impulse

(1 - ηD) Equaliser

Figure 5.6: *Fixed Linear Equaliser (Differentiation)*

noise data was brought back to mainframe computer of Department of Systems & Computer Engineering, Carleton University and reconstructed from the digitised samples. The chapter 4 dealt with the analysis and modelling of this impulse noise data.

Having analysed the impulse noise measurements data captured in a PBX environment, we want to study the effect of impulse noise on a 2B1Q based 2-wire loop ISDN basic rate access system. This effort will not only verify the usefulness of having an impulse noise database but also help a digital system designer in evaluating a ISDN system performance in an impulse noise environment.

The main emphasis of this simulation will be on finding the distribution of the errors per impulse noise events on all the fifteen Bellcore loops (see Table 5.1 for the loop configurations). The impulse noise block (file) consists of all the impulsive events having amplitudes greater than 5mV and also poisson distributed in time with measured mean arrival time. The following few sections will deal with this in detail and the results will also be presented at the end of this chapter.

5.2.1 Simulation Flow

The digital system used for the purpose of evaluation is shown in simple blocks of Figure 5.1 and the functioning of some of its important blocks has also been explained in the section 5.1. In this section, we will explain various characteristics and parameters that were carefully chosen to evaluate the ISDN basic rate access system in an impulse noise environment.

To obtain the performance of the DSL system, the *error statistics* were measured by transmitting 5000 symbols (or bauds) over the fifteen different subscriber loop channels. Following are the parameters used in this system design.

1. Bit rates = 160kbits/sec and 800kbits/sec.
2. Symbol Rates = 80kbaud and 400kbaud (2B1Q transmission).
3. The 2B1Q coding was used here which is nothing but 4-level PAM signalling.

Figure 5.7(a) shows a 2B1Q coder example.

Table 5.1: Bellcore Loop-Plant for testing received signal performance[3]

Loop#	Length (ft)	Description
1.	18000	16500x26,1500x24
2.	16500	13500x26,3000x24,bt1500x24
3.	16000	7500x26,6000x24,bt500x24,1500x24,bt1000x22,1000x22 bt1500x24,bt500x24,btt1000x24
4.	17000	7500x26,4500x24,2000x22,3000x26
5.	15000	9000x26,6000x24
6.	17500	4500x26,12000x24,bt500x24,1000x24,bt500x24
7.	13500	13500x26
8.	16000	9000x24,1000x22,6000x26,bt1000x24
9.	10500	3000x26,bt1500x26,6000x26,bt1500x26,1500x26,bt1500x26
10.	16500	6000x26,3000x22,6000x24,1500x22,bt1000x22
11.	12000	12000x26,bt1500x26
12.	13500	7500x26,4500x24,1500x26
13.	12000	9000x26,2000x24,bt1500x26,500x24,bt1500x26,500x24
14.	14000	3000x26,2000x22,7500x24,bt1000x24,bt500x24,btt1500x26 1500x24,bt1000x24
15.	12000	12000x26

12000x26 => Line length is 12000ft, gauge is 26AWG.; bt => Bridge Tap ;

btt => Bridge tap of a bridge tap ; 1000ft => 0.3048 Km ;

Line specifications are read from left (CO end) to right (Customer end).

Reference: T1D1/87-246, 1987

Please see Appendix for detailed configuration.

4. Transmitted pulses had very low duty cycles. Every pulse (1 sample) is followed by another 7 zero-samples in each baud interval. Alternatively, 8 samples/ baud-interval. See Figure 5.7(b) for an instance.
5. Subscriber loop configurations were the same as tabulated in the Table 5.1 and also shown in Appendix-A. These test loops were found to be working satisfactorily (enough received signal-power) for the basic rate access in the presence of only gaussian noise according to ANSI [3].
6. In this simulation, the noise added at the end of the channel consisted of Gaussian plus Impulse noise. The inclusion of impulsive noise will be described in the next sub-section.
7. The receive filter consisted of two blocks. They are a 4th order, low-pass Butterworth filter and a $(1-\eta D)$ filter to cancel precursor and avoid error propagation in the DFE which follows this. The differentiator coefficient, η used here was equal to 0.8.
8. The output of the $(1-\eta D)$ filter was fed to a sampler which samples the received signal at an instant where the height of the received pulse is maximum. By trial an optimum receiver sampling phase was found for each loop.

The decision feedback equaliser used here is of an adaptive nature. Before starting to look at the system performance using an adaptive DFE, we used a fixed DFE with only gaussian noise as a disturbance, to make sure that the errors caused at the receiver are not due to the adaptation of the DFE. Once being sure that the errors are not being propagated due to the DFE, we used an adaptive DFE with a training sequence of length 800 and disturbance to be gaussian + impulsive noise. The adaptive algorithm used to update the DFE tap coefficients was the LMS (least mean square) algorithm. Some of the parameters used in the DFE are enumerated below:

1. The number of taps were varied from 10 to 25 and 25 taps were found to give satisfactory system performance for these simulations.
2. Training length for the channel estimation was taken to be 800 symbols.

2B1Q Coding (4 Level Pulse Amplitude Modulation)

(a)

Low duty cycle impulse (8 samples/ baud-period)

(b)

Figure 5.7: (a) 2B1Q Coding, (b) Low duty cycle impulse used as a transmitted bit

3. In order to update the DFE tap coefficients, the step size, γ_1 was taken to be 0.005 and this was used during the fast convergence of DFE. The length of fast convergence was taken to be 200 symbols. Another step size, β which was taken to be 0.0001 was used after the DFE had converged. The choice of too large or too small β and γ_1 caused DFE to go out of convergence or to need more symbols to achieve convergence. This choice of β and γ_1 was optimum for DFE to converge properly.

To calculate the signal-to-noise ratio (noise here consisted of gaussian and gaussian + impulsive) the system of Figure 5.1 was used. The 4-level PAM signal was passed through a Code to Voltage level convertor block at a transmission rates of 80kbaud and 400kbaud per second. The function of code to voltage conversion block was to calibrate the transmitted signal to the ISDN standards [3] (also given in Chapter 1 of this work). The calibrated signal was then passed through a selected channel (one channel per simulation). Once again, these channels are taken one by one from the Table 5.1. The output of the digital subscriber loop (DSL) was then filtered using a 4th order, low-pass Butterworth filter. The low-pass filtered noise is the next thing which meets the signal at this point, then the signal and noise are received by $(1-\eta D)$ filter (differentiator with $\eta = 0.8$). This was used to minimize the effect of pre-cursor on the errors in the receiver. The received bits (symbols) are then sampled at every baud-interval and also passed through a DFE. The decision feedback equaliser cancels the maximum ISI within its span. The initial values of tap coefficients were taken to be 0.0. These values were later estimated by the adaptive algorithm depending upon the DSL channel being used.

The calculation of SNR and Probability of errors due to individual impulses is described below.

5.2.1.1 Signal to Noise Ratios

The gaussian noise was passed through a 4th order, low-pass Butterworth filter. This noise is added to the signal received at the output of LP filter. Before adding the

received signal and noise, we calculate the signal and noise power separately (at the output of the LP filter). The ratio of these gave us SNR based on only gaussian noise.

The impulse noise was then added, and the resultant was passed through the LP filter (same as defined above) and then mixed with the received signal. In order to include the impulsive noise with the gaussian noise, a scheme was generated to produce poisson distributed arrival times of impulse events with the measured intra-impulse arrival time. At these times, the algorithm retrieves an impulsive event from the database and adds it to already present gaussian noise. A model was suggested in chapter 1 which is analogous to a ON-OFF switch being controlled by poisson time stampings. This model is reproduced in the Figure 5.8 for the better understanding of a complete system being affected by the poisson distributed impulsive noise and the ever present gaussian noise. The switch S_2 should always be closed and the switch S_1 is controlled by the poisson time stamps. The noise power is once again calculated at the LPF output and SNR was calculated using the same signal power as computed above.

The Figure 5.8 is highly modular in a sense that the same system could be used with gaussian only as well as gaussian + impulsive noise SNR calculation.

5.2.1.2 Error pattern characteristics due to Individual Impulses

The same system of Figure 5.8 could also be used for error measurement with an additional error counter in the receiver and only one LPF. The modified system of Figure 5.9 was used to measure error statistics due to individual impulses.

In the receiver, after baud-interval sampling of the received signal, the signal is detected by the quantiser and then sent to the feedback loop for ISI cancellation. If the estimated and transmitted data symbols are not the same, an error is said to be caused. The error counter increments by one every time an error has occurred.

The scheme flows like this; first the switch S_2 is kept open and the receiver parameters are optimised to get best performance in the presence of only gaussian noise. By doing this we make sure that the channel is being estimated correctly. Then we close the switch S_2 and let the impulsive noise enter too but at exponentially distributed

Simulation Block to calculate Signal to Gaussian+ Impulsive Noise Ratio
on a given DSL Channel

Figure 5.8: DSL System used to calculate SNR where noise is gaussian + impulsive.

time intervals. At this time, we monitor the number of errors being counted every 50 symbols ($\equiv 50 \times 8 = 400$ samples) cumulatively. This enables us to see how the errors are propagating before and after the comparatively higher amplitude impulses hit. It also tells us whether the DFE is converging within its span or not. By synchronising the switches S_1 and S_3 , the Impulsive Errors counter (see Figure 5.9) counts the errors due to each impulse.

The results of above simulations will be presented in the following section.

5.3 Simulation Results

After having described the methodology and various parameters of the DSL system for the calculation of SNR and Probability of errors due to individual impulses, we present the results obtained from the simulations.

5.3.1 Distribution of Number of Errors due to individual impulses

The system of Figure 5.9 was used to compute the errors caused by individual impulses. From the impulsive events database, only 60 events were selected at random for the performance study. While injecting the impulses, it was made sure that their amplitudes were above 5mV. The errors, caused only by the impulses were stored separately for each loop (Loop#1 through Loop#15 of Bellcore).

In order to calculate the distribution of errors due to each impulse, we used the following method to present the results graphically. A mathematical formula is given by,

$$Prob(errors = n \mid Impulse) = \frac{(No. of impulses causing n errors)}{(Total number of impulses in the ensemble)} \times 100\% \quad (5.8)$$

where, n = Number of errors. By using this formula we find the conditional distribution of the number of errors given an Impulse occurs.

The x-axis is divided into seven bins, from 0 to 35 errors. If the first impulse causes 7 errors (say), the counter in the second bin (5 to 10) bin will be incremented by one. If the second impulse causes 1 error (say), the counter of bin 0 to 5 will be incremented

Simulation Block to calculate Probability of Errors due to individual impulses on a given DSL Channel

Figure 5.9: System used for Probability of Errors due to individual impulses computation

by 1. Similarly, the simulation does this for all the impulses. Finally, counters of each bin are counted and divided by the total number of impulses used from the database. This will give the probability that the number of errors equals bin-interval, given that an impulse occurs (see equation 5.8). Alternatively, we can also state this as Probability of error equals to abscissa given an impulse occurs over the ensemble of impulses.

The same simulations were repeated for all the 15 Bellcore loops and the results were plotted to see the distribution of each loop. Using these results one could find bit-error-rate for a loop for which mean impulse arrival time statistics is known. For example, in our case, if we try to find out the BER, it will be very poor. The reason for this could be smaller loop length used for our measurements (maximum 300ft within the building) and the occurrence of impulses was very often at threshold levels of our interests (0 to 35mV). These threshold levels may not be critical in a PBX environment as we have enough received signal strength on the short loops. Whereas, if we try to find out arrival-time statistics for a longer central-office loop, the probability of high amplitude impulses occurrence may not be that high. This may be true because the disturbing impulse amplitudes will be attenuated over the long runs of loops.

Figures 5.10 through 5.17 show plots of the distribution of number of errors. The probability of the number of errors equals abscissa given an impulse occurs (in %) are plotted on the y-axis against number of errors being caused by each impulse on x-axis (error counts). These errors are for 80kbaud transmission rate on different lengths and configurations of two-wire copper loops. By looking at Figure 5.10, we observe that for Loop # 1 (length 18000ft), whenever an impulse hits, there is a 41.5% probability that the number of symbols in error would be between 25 to 30. Whereas, for the same Loop, there are 2% chances that the number of errors would be between 5 to 10 whenever there is an impulsive hit on the line. Therefore, we could consider the first case as a worst case for Loop # 1. Similarly, for other loops, we can see what is the worst probability of errors having occurred whenever an impulse strikes. For instance, for the Loop#2, (see Figure 5.11) there are 24.53% chances that 15 to 20 symbols will be in error whenever an impulse occurs. By observing the other Figures of Loop#3

through Loop#15 (except Loop#5, Loop#8 and Loop#10), we observe that the most prevalent number of errors is between 15 to 20 for 80kbaud system and their chances (probabilities) vary from 24.53% to 35.85% whenever there is an impulsive hit.

The Figures 5.10 through 5.17 also give information about mean and median of the number of errors caused on each loop by the ensemble of impulsive events. These could be displayed in Tables 5.2 and 5.3 for 160kbits and 800kbits per second transmissions, respectively. If we notice these tables closely, we can easily derive an average estimate of numbers of errors being caused by the impulsive events. For the better estimations of number of errors due individual events statistics, the medians are also displayed in these tables.

Figures 5.18 through 5.25 display the corresponding distributions for 400kbaud transmission. By observing these as well as the tables of mean and median number of errors, we find that usually impulse noise events generate fewer errors in the 400kbaud system, presumably because of less spreading of the noise waveform by the higher bandwidth filter.

Table 5.2: *Mean and Median of Number of Errors for all the 15 loops at 160kbps/sec.*

Loop No.	Mean of Number of errors	Median of Number of errors
1	31.6	28
2	19.1	19
3	18.4	17.5
4	18.7	18
5	16.5	15.5
6	18.2	17
7	15.9	15.5
8	17.1	16
9	15.2	15.5
10	17.3	16
11	15.4	15.5
12	15.1	15.5
13	15.2	15.5
14	15.6	15.5
15	15.2	15.5

Table 5.3: *Mean and Median of Number of Errors for all the 15 loops at 800kbts/sec.*

Loop No.	Mean of Number of errors	Median of Number of errors
1	15.5	12
2	9.5	4
3	11.2	10
4	14.1	11
5	12.8	11
6	14.1	12
7	12.2	11
8	12.7	11
9	10.7	10
10	12.3	10
11	11.3	10
12	17	11
13	11.3	10
14	11.2	10
15	9.5	9

Figure 5.10: *Probability of Errors Vs. Number of Errors for Loop #1 (80kbaud)*

Figure 5.11: Probability of Errors Vs. Number of Errors for Loop # 2 & 3(80kbaud)

Figure 5.12: *Probability of Errors Vs. Number of Errors for Loop # 4 & 5(80kbaud)*

Figure 5.13: Probability of Errors Vs. Number of Errors for Loop # 6 & 7(80kbaud)

Figure 5.14: *Probability of Errors Vs. Number of Errors for Loop # 8 & 9(80kbaud)*

Figure 5.15: Probability of Errors Vs. Number of Errors for Loop # 10 & 11(80kbaud)

Figure 5.16: Probability of Errors Vs. Number of Errors for Loop # 12 & 13(80kbaud)

Figure 5.17: Probability of Errors Vs. Number of Errors for Loop # 14 & 15(80kbaud)

Figure 5.18: *Probability of Errors Vs. Number of Errors for Loop # 1 (400kbaud)*

Figure 5.19: *Probability of Errors Vs. Number of Errors for Loop # 2 & 3(400kbaud)*

Figure 5.20: Probability of Errors Vs. Number of Errors for Loop # 4 & 5(400kbaud)

Figure 5.21: Probability of Errors Vs. Number of Errors for Loop # 6 & 7(400kbaud)

Figure 5.22: Probability of Errors Vs. Number of Errors for Loop # 8 & 9(400kbaud)

Figure 5.23: Probability of Errors Vs. Number of Errors for Loop # 10 & 11(400kbaud)

Figure 5.24: Probability of Errors Vs. Number of Errors for Loop # 12 & 13(400kbaud)

Figure 5.25: *Probability of Errors Vs. Number of Errors for Loop # 14 & 15(400kbaud)*

Chapter 6

Conclusion & Suggestions

6.1 Conclusion

The growing demand for voice and data integration over the subscriber loops has prompted the development and enhancement of high speed data transmission. The theme of this research was to *characterise the real-time measured impulse noise data on two-wire copper loops in a business PBX loop-plant environment, and assess the performance of a digital subscriber loop (DSL) system for envisaged high bit-rates (160kbps & 1.544Mbps) in an impulsive noise environment.* The main emphasis in this work has been given to the characterisation of impulse noise on subscriber lines connected to a PBX, in contrast to other studies for central office environments.

As we have seen the structure of this work, the research started by designing measuring equipment. The commercially available hardware and software was tailored to serve our purpose of real-time measurement of impulse noise. The measuring equipment was first tested in a laboratory environment by generating the sharp impulses using impulse generator as well as telephone cradle switch (the pulses were fed through a current-probe in this case). The bandwidth of measurement and the detection levels were also varied to check the proper functioning of the controlling software and the IEEE bus. The maximum bandwidth of the digitiser was 50MHz which was found to be very suitable for these impulsive measurements.

An approach to finding a typical impulse which represents an ensemble of impulse noise data would help a digital system designer in designing a receiver to combat

impulsive nature of disturbances. Keeping this in mind, the shapes of impulsive events, which were highly random have been classified in three different classes and a typical pulse for each class has been displayed (see Figures 4.41 through 4.43). Since the shapes of the events in each class were very diverse, it was not possible to come-up with a single typical impulse for the ensemble.

This was followed by frequency domain analysis to see the band of frequency within which impulse noise spectra usually lies. From the measurements it was found that impulse noise spectra lies in 0 to 250kHz region and the energy spectrum was fairly flat within 0 to 50kHz (the 3dB filter cutoff for 160kbps system was 40kHz) which was in contrast to crosstalk power-spectra which increases with frequency.

The various impulse arrival-times have been presented at four threshold levels (5, 10, 20 and 35mV). Along with this, it was also investigated what percentage of impulse events (at a particular threshold of detection) arrive at different times of the day. This was very steady during week days and most of the activities were seen to occur before the lunch time (12 O'clock). This also reflects the human behaviour in using the communication equipment in an automated office-environment.

The conditional distributions of amplitude and duration of impulsive events were also computed from the re-constructed impulsive noise data. The conditional amplitude and duration distributions were plotted with respect to the receiver threshold levels. With these distributions, we thought it would be more helpful if we had provided the dependence of number of positive crossings on their detection levels. This is presented in the form of a plot in Chapter 4.

Finally, to assess the performance and the behaviour of a DSL system, the author saw a need to simulate a conventional digital subscriber loop system and then study that in an impulsive noise environment. The Chapter 5 as a whole has been dedicated to this study. We used the 15 channels provided by Bellcore as standard loops to assess the received signal performance (see Appendix-A for the loop configurations). The conditional distribution of the number of errors given that an impulse occurs have been displayed for all the 15 loops at 160kbps (80kbaud) and 800kbps(400kbaud) bit-rates.

These results project a picture of number of errors being caused given an impulse strikes. Looking at these results more closely and relating them with the loop configurations (shown in Appendix A), one could see how the loop performance varies with length, wire-gauge and bridged-taps along the length of the loops. For instance if we look at the results for Loop#7 and Loop#8 (see Figures 5.16 and ?) for 80kbaud transmission, we notice here that the probability that the worst-case errors will be between 10 to 15 has increased with the increased complexity in Loop#8 (longer loop with different gauges and a bridge-tap). The same phenomenon can also be noticed if we compare the results of Loop#11 and Loop#12 (see Figures ? and ?) at 80kbaud. In this case also, the percentage of errors has increased with the addition of a bridge-tap in Loop#11.

The Tables 5.2 and 5.3 presented in the Chapter 5 display the mean and median of number of errors due to individual impulses for all the 15 loops at the 160kbits/sec and 800kbits/sec transmission rates, respectively. For the 160kbits/sec transmission rate (see Table 5.2), most of the loops show mean number of errors between 15 and 19 whereas the median statistics for the same varies from 15.5 to 18 for most of the loops.

On the other hand, for 800kbits/sec. transmission rate, the mean number of errors for most of the loops vary between 11 to 16.98 whereas the median of number of errors varies between 10 to 12. The mean and medians statistics for 800kbits/sec system are relatively lower (see Table 5.3) than the 160kbits/sec system statistics. This could presumably be because of lesser spreading of the noise waveform by the higher bandwidth filter.

It is also seen that the probability of errors increases by increasing the transmission rate.

6.2 Suggestions for Future work

During the course of these measurements, the author has few suggestions which will help greatly for this kind of measurement and analysis study in future.

1. While analysing the re-constructed impulse noise data obtained from the PBX loop plant of MITEL Corporation, Kanata, it was realised that the intra-impulse

arrival time statistics was concentrated in the interval 1.3 to 1.4 seconds for 5 and 10mV threshold of detection. It appears that because of limitations imposed by the clock-rate on the personal computer and the IEEE bus, most of the impulses were gathered around this time stamping. It is suggested here that for more realistic intra-arrival time statistics hardware would have to be developed which makes use of assembler and interfaces with GPIB bus and also has flexibility to change measurement bandwidth and time-settings.

2. It was also felt that more intra-arrival time measurements have to be carried out to get a better estimate and distribution of time of arrival. Future research should dedicate a complete measurement session to the impulse intra-arrival time statistics.
3. The use of $(1-\eta D)$ filter before the adaptive equaliser has improved the error performance in the system considered in this work. As suggested by reference [21], a further investigation could be done to make use of nonlinear C-filters for impulsive noise affected receivers to enhance the receiver performance.
4. The simulations carried out in this thesis (in impulsive noise environment) using an adaptive DFE receiver should be extended to Trellis Coded receiver to see if this outperforms the DFE in impulsive noise environment.

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Appendix A

Test Subscriber Loops

For testing the received signal performance, the following 15 loops had been considered through out this work. See Figures 1 through 3. These loop configurations have been contributed by Bell Communications Research Laboratories (Bellcore) towards the standardisation of received signal performance evaluation. The ANSI (American National Standards Institute) has also considered these loops as standards for performance assessment of high bit-rate 2B1Q based system.

The each loop has one hybrid connected on each end of it. The CO end marked on the loops is the Central Office and other end is the Customer end. The data is transmitted from both sides. The horizontal lines are the length of the loop whereas the vertical lines are the bridge-taps. Three different gauges have been used here, these are namely 22, 24 and 26 AWG (American Wire Gauge). The loop lengths are marked in feet²⁹.

²⁹ 1000 feet = 0.3048 Km.

Figure-A 1: Loops for Testing Received Signal Performance (Loop # 1 to Loop # 5)

Figure-A 2: Loops for Testing Received Signal Performance (Loop # 6 to Loop # 10) (Contd.)

Figure-A 3: *Loops for Testing Received Signal Performance (Loop # 11 to Loop # 15) (Contd.)*

Appendix B

Pseudo-Code Listing for Line IR Calculation

In order to compute the impulse response (IR) of the 15 loops (mentioned in Appendix-A), a code was generated on Sun-OS4. The transmission rates, 80kbps 400kbps are considered through out the simulation. Section 5.1.2 explains the calculation of A, B, C, D line parameters. The cable characteristics were taken from the reference [45]. These frequency characteristics of 22 AWG, 24 AWG and 26 AWG cables at 70°F include $R(\Omega/\text{mile})$, $L(\mu\text{H}/\text{mile})$, $C(\mu\text{F}/\text{mile})$, $Z_0(\Omega)$, $G(\text{mho}/\text{mile})$ and $\gamma(\text{dB}/\text{mile})$ at frequencies varying from 1Hz to 5MHz.

Pseudo-Code for IR Calculation:

Open the files containing Cable characteristics.

```
Read the parameters in the arrays.
#define sampling_frequency 3.2e6
#define number_of_samples 8192
Also define hybrid parameters, Zb,R,C and L
```

Calculate the ABCD parameters.

```
for(i=2; i<= number_of_samples; i++)
{
    m = hyb_lo(i,xifreq(i),L,C);
/* Loop-to output in hybrid */
    m = line(i,1500,gamma24,z24)*m;
/* For a section of line of length 1500ft, 24AWG. */
    m = bt(i,500,gamma24,z24)*m;
/* For a bridge-tap of length 500ft,24AWG */
    m = hyb_tl(i,xifreq,R,L,C,Zb)*m;
/* Transmitter-to-loop hybrid */

    a = m[1,1];
    b = m[1,2];
    c = m[2,1];
```

```
d = m[2,2];
```

```
}
```

Calculate the frequency response.

Window it using Blackman window.

Take IFFT of this response to get impulse response.

Write it on a file in columnar format.

**This will be read by the main simulation as an
impulse-response.**

Do this for all the 15 loops at both the transmission-rates.

END

1 3|0 9|9|0

FIN