Telephone Line Data Transmission Using LocalTalk

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By

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Telephone Line Data Transmission Using LocalTalk

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Abstract

With the growth of the Internet, more and more people want their computers to be connected to the Internet and to get information from the Internet as fast as possible. Presently available remote network access methods are limited in speed and cannot support simultaneous voice transmission. They do not fully satisfy the Internet users' needs.

This thesis presents a new high speed network access technology. With this method, a residential telephone line can be used to connect home computers at 230 kbits/s to the Internet without loss of the voice service. In order to put this system in practical use, some problems, such as the interference between the voice and data, the limited transmission distance and the effect of a bridged tap on the telephone line, must be solved. These problems and their solutions have been discussed in this thesis.

Acknowledgments

I wish to express my special thanks to my supervisor Professor David E. Dodds. I would like to thank him for his helpful guidance and wisdom during the course of this work.

My work would not have been possible without the outstanding help from Greg Erker and Garth Wells, the research engineers in TRLabs Saskatoon. I would like to thank Cliff Klein, the director of TRLabs Saskatoon, for the financial assistance, the first class equipment and the excellent environment that TRLabs provided during this work. The help that Leon Lipoth, Denard Lynch, Jack Hanson and others at TRLabs gave me is greatly appreciated. I also would like to thank Trudy Gall.

Finally, I want to thank my wife, Yanping, my parents and my sister for their consistent encouragement and support.

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List of Abbreviations

ACK	Acknowledgment
ADSL	Asymmetric Digital Subscriber Lines
AWG	America Wire Gauge
CATV	Community Antenna TV (Cable TV)
CCITT	International Telegraph and Telephony Consultative Committee
CSMA/CA	Carrier Sense Multiple Access with Collision Avoidance
CTS	Clear-to-Send
CRC	Cyclic Redundancy Checking
dB	Decibel
dc	Direct Current
dBm	Decibels above One Milliwatt
dBmC	Decibels above C-messaged Reference Noise
DDP	Datagram Delivery Protocol
DPLL	Digital Phase Locked Loop
DSP	Digital Signal Processing
DTMF	Dual Tone Multifrequency
ENQ	Enquiry
EPL	Equivalent Peak Level
HDSL	High-bit-rate Digital Subscriber Lines
IDG	Inter-Dialog Gap
IFG	Interfame gap

ISDN	Integrated Services Digital Network
ISI	Intersymbol Interference
ISP	Internet Service Provider
JWI	Jumper Wire Interface
FM-0	Biphase Space
kbits/s	Kilobits per Second
LLAP	LocalTalk Link Access Protocol
Mac	Macintosh
Mbits/s	Megabits per Second
NRZ	Nonreturn to Zero
NRZ-L	Nonreturn to Zero Level
NRZ-S	Nonreturn to Zero Space
NRZ-M	Nonreturn to Zero Mark
PBX	Private Branch Exchange
PC	Personal Computer
PIC	Polyethylene Insulated Cable
PN	Pseudo-noise
PCM	Pulse Code Modulation
R-C	Resistor - Capacitor
RTS	Request-to-Send
RZ	Return to Zero
SCC	Serial Communications Controller
S/N	Signal to Noise ratio
VU	Volume Unit

Chapter 1: Introduction

With the extension of the Internet, more and more people are able to access this network from their computers at home. One study shows that the Internet subscriptions will increase up to 14 million by the end of this century [1]. On the other hand, the information in the Internet grows larger and larger. On-line CD quality music, full motion video and high resolution pictures are available on the Internet. However, these kinds of data need a high speed access method to download. With the increasing demand for higher speed on-line access, how to increase the Internet access speed to the home and keep the cost low becomes one of the hot topics in network access studies.

1.1 Today's Network Access Methods

In order to reduce the cost, today's network access methods for the home are all based on the already existing wire plant, such as subscriber lines in telephone networks and cables in cable television networks.

The most popular and cheapest way to access the network information is realized by a voiceband modem. Fig. 1.1 illustrates the network access via a modem.

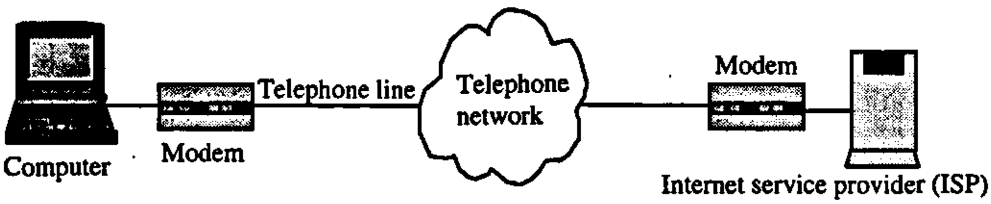

Fig. 1.1 Network access by a voiceband modem

The computer is connected to the modem and the modem is connected to the telephone network via a copper wire pair. When the computer transmits data, the modem modulates the data to analog signals whose characteristics are suitable to telephone network transmission. When receiving data, the modem needs to demodulate the analog signals which come from the telephone network and to send the demodulated data to the computer. On the other end, the Internet service provider (ISP) also needs a similar modem to modulate or demodulate the data signals. Through the pair of modems, the computer can dial up to ISP and access the Internet. The advantages of this method are convenient and cheap. A user can access the Internet where a telephone is available. The price of a modem is less than \$150 and monthly charge by the ISP is about \$20. However, the throughput of the voice band modem is low. The maximum available data rate for a voiceband modem is about 33.6 kbits/s and the speed will drop down if the noise of the telephone line is large. If a modem is used, voice service can not be offered at the same time on the telephone line. In addition, the Internet access via a voice modem causes impact to the telephone networks as well. Since the modem users generate a lot traffic and occupy excursive resources in the telephone networks, the service quality of the telephone network decreases with the increase of Internet users.

The second commercial method which is available for Internet access is through basic rate ISDN (Integrated Services Digital Network). ISDN attempts to combine various telecommunications services, such as voice, fax and binary file transfer, into a unified system [2]. ISDN needs digital switching systems and ISDN subscriber lines. The ISDN Internet access is similar to Fig. 1.1, but the voiceband modems at the computer side and the ISP side are substituted by a pair of ISDN terminal adapters (or ISDN modems) that can support ISDN protocols. An ISDN modem can support two 64 kbits/s channels. A user can use one 64 kbits/s channel as a voice channel and another one as a data channel or both channels as data channels. Therefore, ISDN can support voice and 64 kbits/s data simultaneous transmission or exclusive 128 kbits/s data transmission. Obviously, ISDN access is faster than voice modem access, but the price for ISDN is higher. An ISDN modem costs about $300 and the rent for ISDN subscriber line from the local telephone company is about $80 per month. The ISDN installation requires careful integration of the telephone company service, the terminal adapter, the computer and the software, which can not be conducted without knowledge of ISDN. ISDN users still need to dial the ISP through the central office, therefore the switch resources are still needed in ISDN calls.

Other high speed Internet access methods via the telephone copper wires, such as HDSL (high-bit-rate digital subscriber lines) and ADSL (asymmetric digital subscriber lines) are under development and have been put into commercial service on a limited basis. HDSL is a scheme that initially will use twisted pairs to transport an unrepeated symmetric duplex data within 3.2 km [3]. ADSL is another high speed access scheme that supports 1.544 Mbits/s downstream (to customer) and as much as 640 kbits/s upstream data transmission [4]. HDSL and ADSL are high speed Internet access methods and are considered as transition technologies to optical fiber network access in the future [5]. However, the cost of HDSL and ADSL are still very high and no commercial mass deployment is available. According to the available information from the HDSL and ADSL trials, a HDSL or ADSL modem at the customer side costs about $1000 to $2500 and HDSL or ADSL users will be charged at $60 to $100 per month for Internet access [6].

Another high speed Internet access method, which is a competitor to ADSL, is the cable modem. A cable modem is a device that allows high speed Internet access via a cable TV (CATV) network [7]. A cable modem can deliver 10 to 30 Mbits/s downstream data. However, this bandwidth is shared by hundreds of customers who are connecting to the same cable segment. Therefore, the actual speed that a user can achieve varies with the traffic and the number of users connecting on the cable. The speed of the upstream channel is in the range of 500 kbits/s and this channel is also shared by all users on the cable segment. Since today's cable network is built to provide television broadcast services, it can support broadcast communications very well but not two-way communications. Cable companies have to invest money to upgrade their networks to two-way systems. Another critical problem of the cable network is that most business offices are not connected to cable TV networks. This disadvantage prevents cable modems being used in business offices. For home users, the price of the cable modem is still too high and there has not yet been mass deployment [8].

It is clear that today's network access methods have their advantages and disadvantages. The voice modem is cheap and convenient, but access speed is too slow. Although ISDN has higher data rate, the price is high and the service is not available everywhere. HDSL, ADSL, and cable modem are extremely expensive and are not affordable by today's Internet users.

1.2 A New Network Access Method

Is there any other way that can provide fairly high speed network access with low cost? The answer is positive. A new network access method, which has been developed by Prof. D.E. Dodds, is implemented in this project. The objective of this low cost network access method is to support voice and 230 kbits/s data transmission simultaneously on the existing 26 gauge telephone lines. This work is based on the interface module has been developed by Greg Erker and Garth Wells in TRLabs. The interface module (will be introduced in Chapter 5) has a high pass filter and a low pass filter that can separate the data signal and the voice signal in different frequency bands, therefore, this interface module can realize the data and voice simultaneous transmission. However, the data transmission distance is limited to 700 meters. Therefore, for practical applications, the distance of the data transmission is proposed to be longer than 2 km, which is suitable for the applications in telephone network distribution area (see Chapter 6 for detail), small cities, hotels, hospitals and industrial parks. The Internet access system with this new technology is illustrated in Fig. 1.2.

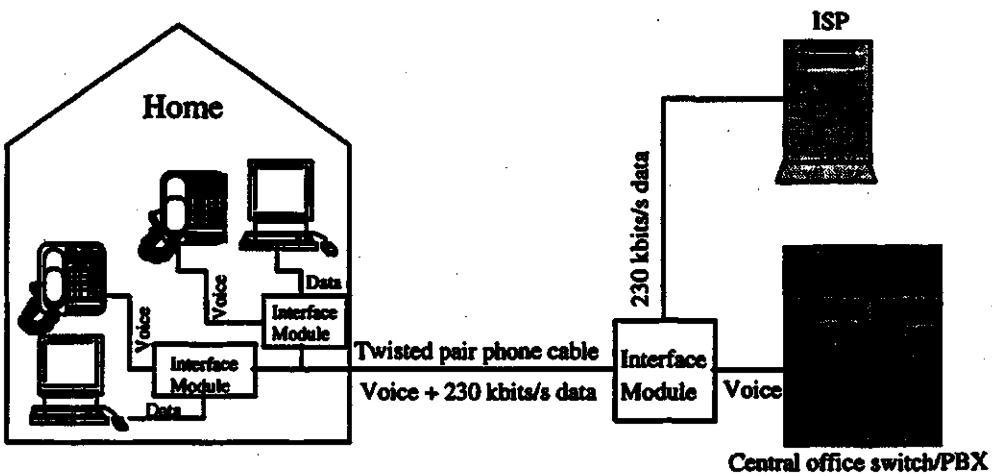

Fig. 1.2 The new network access method

In this network access method, special interface modules are required at the home and central office or PBX (Private Branch Exchange) sides. The interface module can combine/separate the data and voice signal to/from the telephone line without interrupting the services offered by each other. At home, the telephones and the computers are connected to a telephone line via the special interface modules. On the other end, the voice signal from the central office switch (or a PBX) and data from an ISP are also connected to the telephone line via a similar interface module.

This new access technology uses commonly-used LocalTalk protocol (described in Chapter 2). This protocol is built in all Macintosh computers and can be also performed by all IBM PC compatible computers. The protocol is robust and allows multiple computers to be connected to one link. Therefore, this access method is able to support multiple computer network access, which is desirable by small business offices or those that have multiple computers at home. Unlike voice modem or ISDN, the new network access method has a permanent data connection to an ISP and has no impact on the telephone network [9]. The speed of the new network access method is also higher than the voice modem and ISDN (230 kbits/s vs 33.6 kbits/s and 128 kbits/s). Comparing to the Cable modem and ADSL, the new network access method is much cheaper and has a higher function to price ratio. Since the technology used in this project is less complex and no extra wiring is required in the house and the subscriber line, the cost of this network access is low. The proposed price for the interface module is about $100 to $150 [6], which is equal to today's voiceband modem, but the user can have 230 kbits/s permanent data link and simultaneous voice service.

1.3 Thesis Objectives

The objectives of this thesis are to extend the 230 kbits/s data transmission distance longer than 2 km and examine the data distortion caused by a bridged tap.

When the LocalTalk data are transmitted on a telephone cable, the telephone cable will cause distortion to the data and limit the data transmission distance. In order to extend the data transmission distance on the telephone line, we should understand how the telephone cable distorts the LocalTalk data and find a way to compensate for the distortion. Therefore, to understand the transmission line theory, the characteristics of the telephone cable and the data transmission theory can help us to find solutions for extending the data transmission distance.

The bridged taps in the telephone cable cause distortion in the high speed data. We need to investigate the seriousness of the distortion that bridged taps can cause to the data signal. Solutions to bridged tap distortion will be investigated.

1.4 Thesis Overview

In this thesis, Chapter 2 describes some background on AppleTalk computer protocols and LocalTalk local area network. This chapter also examines why LocalTalk is used in this project. Transmission theory and the characteristics of twisted pair cable that is used in telephone subscriber loops are discussed in Chapter 3. Equalization, predistortion and some background on baseband data transmission are presented in Chapter 4. Chapter 5 introduces a key circuit that can isolate the LocalTalk data from voice signals on the subscriber line. With this circuit, data and voice can be transmitted simultaneously on the same phone line without interfering with each other. At the end, Chapter 6 presents the effect of the bridged tap and methods that can be used to extend the data transmission distance. Some experimental data based on laboratory tests is presented in this chapter to support the success of the new technology.

Chapter 2: The LocalTalk Local Area Network

The LocalTalk/AppleTalk local area network system is a set of hardware and software specifications that allows a Macintosh computer to communicate with peripheral devices, servers and other computers [10]. It was first developed by Apple Computer Inc. in 1984 [11]. In June 1989, Apple Computer Inc. introduced AppleTalk phase 2, which is compatible to the 1984 version and functions better in large network environments [12].

Each Macintosh computer can perform AppleTalk protocols and has the standard LocalTalk network interface hardware. The PC computer also can connect to a LocalTalk network via a LocalTalk board. AppleTalk is a network protocol that was originally designed to operate on the LocalTalk physical layer.

2.1 AppleTalk Protocol

As illustrated in Fig. 2.1, AppleTalk is a layered network protocol. The advantage of the layered network protocol is that lower level layers offer transparent services to higher layers, thus the developer can easily access, change or add certain layer protocols without disturbing other layer protocols [13].

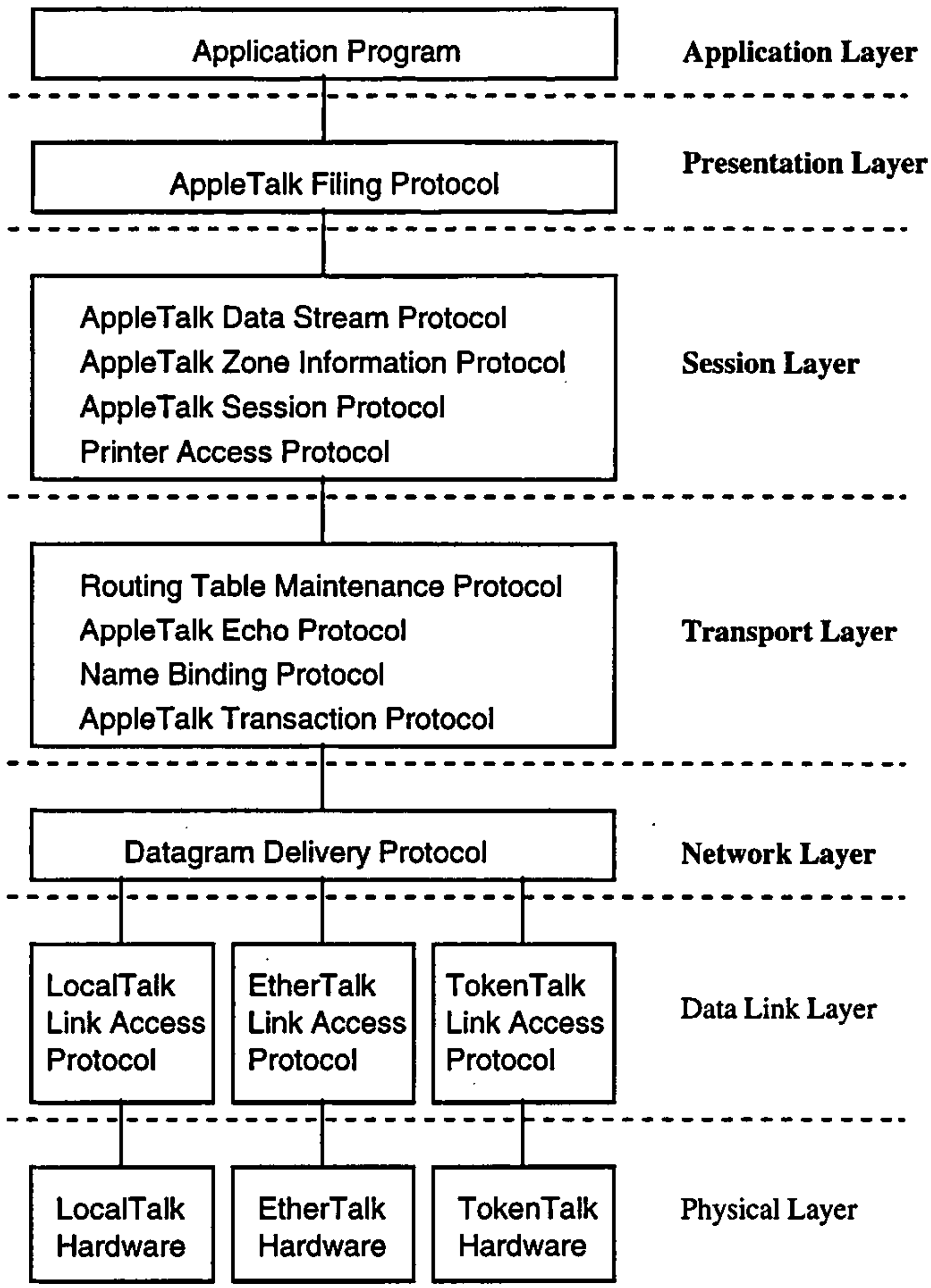

Fig. 2.1 AppleTalk protocol layers

The upper layer AppleTalk (layers in bold) protocols can run on a variety of data transmission media using different connectivity infrastructure; these upper layer AppleTalk protocols are independent of the physical link. Any change in the communication hardware (physical layer) and the associated protocols for controlling access to the hardware links (data link layer) is transparent to the upper layers of AppleTalk protocols.

LocalTalk, EtherTalk and TokenTalk are names of different media access which are used by upper layer AppleTalk protocols. EtherTalk uses standard Ethernet technology to transmit data at 10 Mbits/s. It makes higher layer AppleTalk protocols run on an Ethernet network. TokenTalk provides connection to industry-standard token ring networks. Similar to EtherTalk, TokenTalk can fully support higher layer AppleTalk protocols as well. LocalTalk is a serial cable networking medium that transmits data at 230.4 kbits/s. It has some interesting characteristics:

(1) The LocalTalk's receiver, transmitter and the LocalTalk Link Access protocol are built into every Macintosh computer. An ordinary telephone cable can be used as a LocalTalk common transmission link as well, so the network is easily set up and inexpensive.

(2) 230.4 kbits/s data rate is fairly rapid compared to voice band modems and this high speed data can be transmitted longer than 2 km, the average distance of the telephone network distribution area (detailed in Chapter 6). The LocalTalk Link Access protocol can also function properly within the delay associated with this range (described in Section 2.4.3).

(3) The power spectral density of the LocalTalk signal is suitable for data and voice simultaneous transmission (see Section 2.3.4).

These characteristics make LocalTalk the choice for this project. More details about the LocalTalk network are described in the following sections.

2.2 LocalTalk Network Configurations

A LocalTalk network can be configured as a stand-alone local area network, as illustrated below in Fig. 2.2. This kind of configuration allows computers in the network to exchange information and share resources such as printers and file servers. This group of computing devices is known as a work group.

Fig. 2.2 Stand-alone LocalTalk network

The stand-alone LocalTalk network can be expanded to a larger LocalTalk network by the use of a bridge. There are two kinds of bridges. One is called a local bridge as shown in Fig. 2.3. The local bridge is used to interconnect several LocalTalk networks that are close to each other.

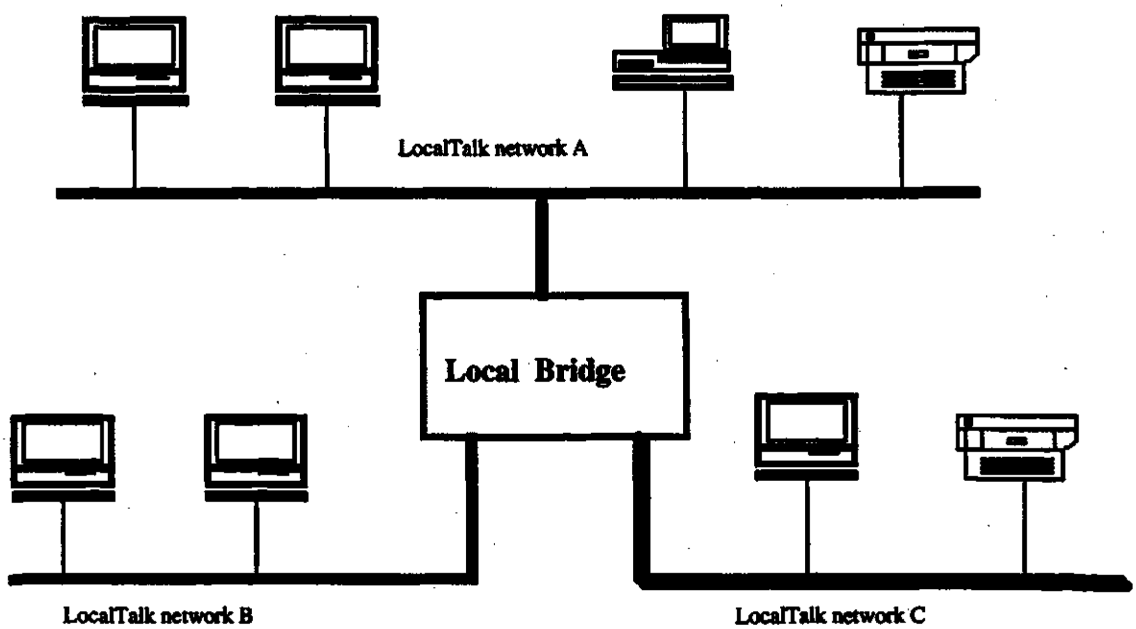

Fig. 2.3 The configuration of a local bridge

Another type of bridge, called a half bridge, is shown in Fig. 2.4. Two half bridges are interconnected by a long distance communication link. Each half bridge is directly connected to a LocalTalk network. The two half bridges and the communication link together have the same function as the local bridge. The advantage of the half bridge configuration is that it can interconnect two or more LocalTalk networks that are far away from each other. But this advantage is attained at the cost of lower throughput and less reliability due to the long distance communication link.

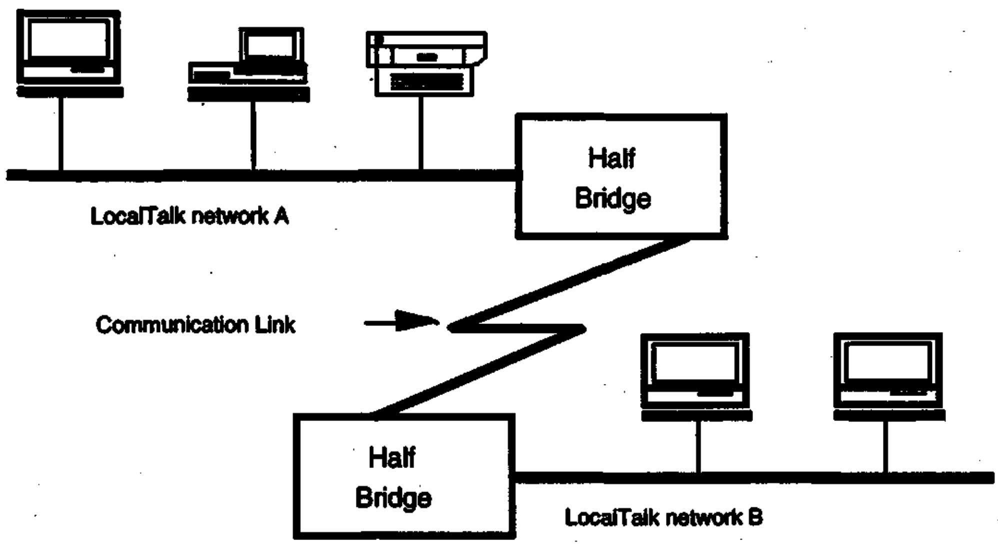

Fig. 2.4 The configuration of a half bridge

Theoretically, a single LocalTalk network can support up to 254 devices and the maximum number of LocalTalk networks that can be interconnected is 65,534 [11]. LocalTalk networks can be interconnected to EtherTalk and TokenTalk networks by an AppleTalk Internet Router and also can be connected to other non-AppleTalk network systems through a gateway [12].

2.3 LocalTalk Physical Layer

LocalTalk physical layer performs the functions of bit encoding, decoding, synchronization, signal transmission, reception and carrier sense. Fig. 2.5 shows the diagram of LocalTalk hardware.

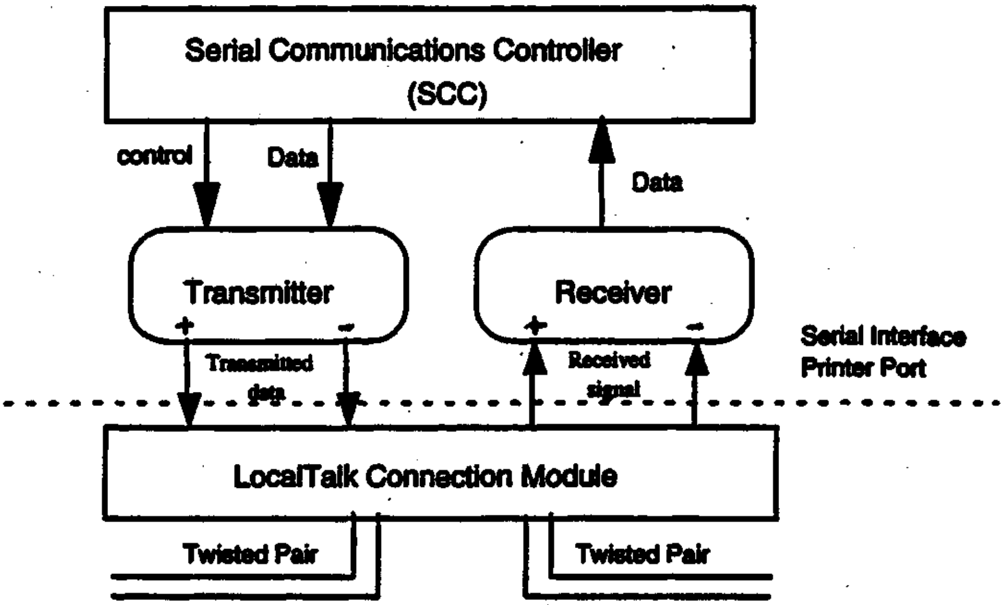

Fig. 2.5 Diagram of the LocalTalk hardware

The hardware of the LocalTalk physical layer is composed of a serial communications controller (SCC), a transmitter, a receiver and a LocalTalk connection module. Data are encoded by the SCC and transmitted to the LocalTalk connection module by the transmitter. On the inverse direction, the receiver receives data from the LocalTalk connection module and sends the data to SCC for decoding. The transmitter, receiver and the SCC are built in the Macintosh computer serial interface. Details of the Macintosh serial interface are shown in Appendix 1 [14]. The LocalTalk connection module is an interface between the computer and the network transmission media (twisted pair). With the LocalTalk connection module, Macintosh computers can be connected in physical daisy chain as shown in Fig. 2.6.

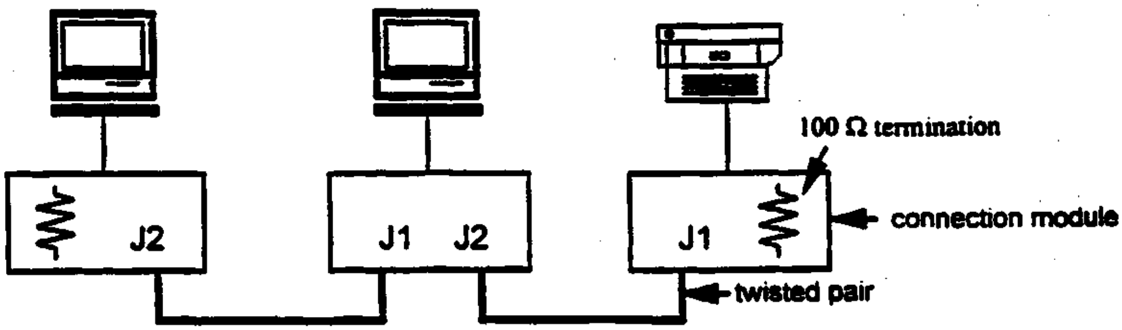

Fig. 2.6 Connection modules in a LocalTalk network

In the following sections, each function block shown in Fig. 2.5 will be examined in detail.

2.3.1 LocalTalk Connection Module

There are two versions of the LocalTalk connection modules. One is the standard LocalTalk Connection Module which connects the Macintosh printer port with shielded twisted pairs. A simplified LocalTalk Connection Module is illustrated in Fig. 2.7. There are two 3-pin miniature DIN connectors between transformer and twisted pair. Each 3-pin connector has a coupled switch that can automatically connect a 100 Ω termination (R2) across the line when only one of the connectors is used. Thus, with the use of the connection module, a device can be removed freely from the network without disturbing the network operation.

R3 and R4 are used to protect the receiver from overcurrent and R1 provides static drain for the cable shield to ground [11].

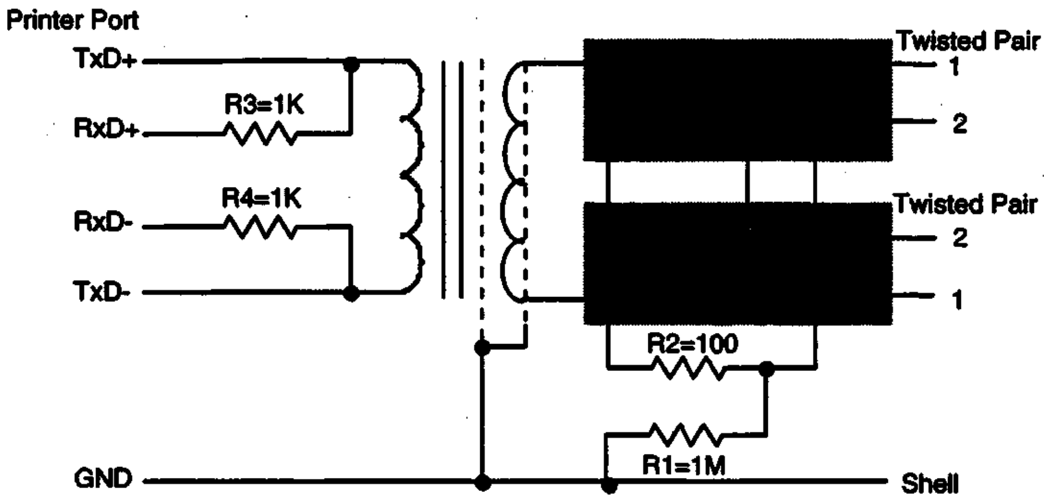

Fig. 2.7 LocalTalk connection module

The shielded twisted pair cable used in LocalTalk has characteristic impedance of 78 Ω. The 100 Ω termination resistor is selected to prevent signal reflection at the ends of the cable. Detail explanations are described in Chapter 3 transmission line theory.

The transformer used in the connection module requires high common mode rejection and sufficient bandwidth. The common mode signals are unwanted signals which are applied to both transformer inputs with the same amplitude and polarity. As shown in Fig. 2.8, V_{i1} and V_{i2} are common mode input signals. For an ideal transformer, the output Vo should be 0 V when the input is a common mode signal. Practically, the transformer can achieve high common mode rejection by using symmetric windings in both the primary and the secondary.

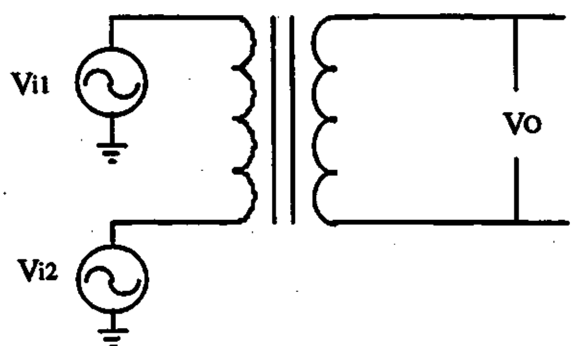

Fig. 2.8 Common mode input signals

In the LocalTalk connection module, the transformer is specified with a lower cutoff frequency at 1.1 kHz and the upper cutoff frequency at 940 kHz, which is sufficient to transform the LocalTalk data [15].

The LocalTalk network using the connection modules of Fig. 2.7 can interconnect up to 32 nodes (computers, printers and servers) within a total cable length of 300 meters [11].

Another version of the connection module is the PhoneNet connector that has been developed by Farallon Computing Inc. This connector uses two RJ-11 phone jacks instead of the 3-pin miniature DIN connectors. The coupled switch does not exist in this connector. The resistor R2 is replaced by an external 120 Ω resistor which is connected manually to terminate the ends of the network. In addition, the PhoneNet connector is designed to use ordinary unshielded 22 to 24 gauge phone cable, It uses the black and yellow wires of a 4-wire phone cable, leaving the red and green wires available for the normal telephone line. The use of PhoneNet connectors and telephone wire increases the total length of the LocalTalk network up to 1000 m [10]. In addition to the series network connection shown in Fig. 2.6, the connectors can be connected in a star topology as illustrated in Fig. 2.9. The total length of all cable segments (such as segment A - B) must be less than 1000 meters. The advantage of this topology is that a PhoneNet connector can be plugged or unplugged from the backbone without affecting other devices.

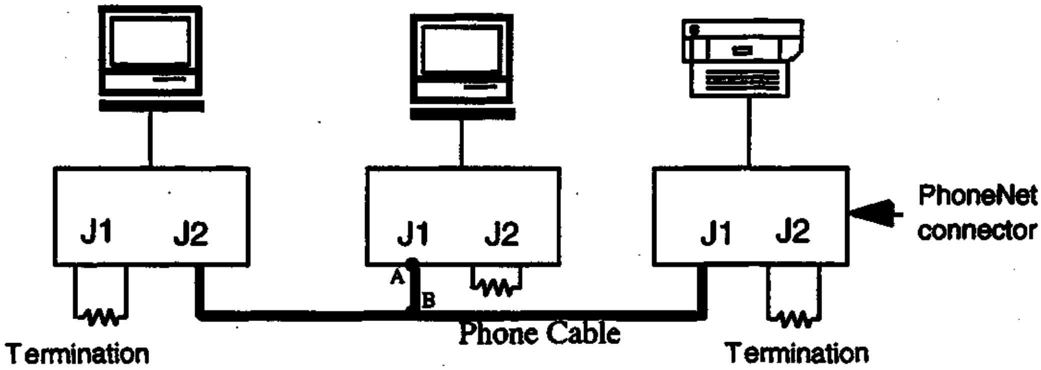

Fig. 2.9 Star topology

2.3.2 Transmitter and Receiver

In older Macintosh computers, AM26LS30 and AM26LS32 integrated circuits were used as the Apple LocalTalk transmitter and receiver. New computers use a single transceiver chip to perform the functions of the AM26LS30 and AM26LS32 chips. The AM26LS30 chip is supplied by both positive and negative voltage of 5 V. It has +5 V, -5 V[1] and high impedance 3-state output [16]. When there is no data transmission, the transmitter is switched to high impedance status (disabled) by the serial communication controller (SCC). The high impedance state prevents idle transmitters from interfering with active ones. The function of the Apple LocalTalk line driver is illustrated in Fig. 2.10.

The unipolar input signal to the transmitter has low level amplitude at 0 V and high level amplitude at 5 V. The transmitter generates two bipolar output signals, A and B. Signal B is nothing but the inverted copy of signal A. They are called balanced signals. If signal B is considered as a reference to signal A, there are ±10 V differences (see point c and d) between them. Thus, we can treat the LocalTalk data as a bipolar signal with ±10 V amplitude.

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1 In practice, due to the structure of the driver, the outputs of the driver are about ±4 V. However, for easier calculation and studying, ±5 V is used in this thesis.

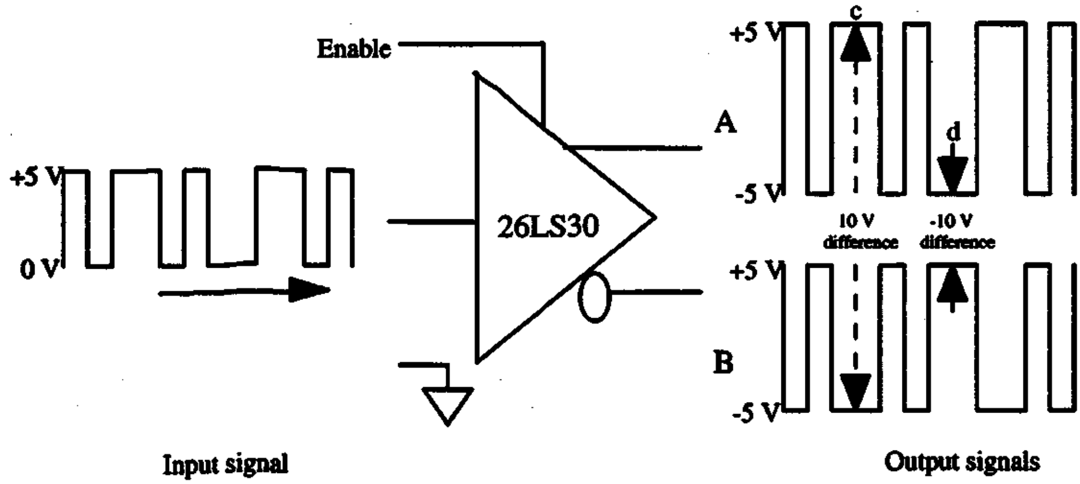


Fig. 2.10 LocalTalk line driver and balanced signals

The LocalTalk differential receiver uses single +5 V power supply. It receives the balanced signals from the connection module and recovers them to an unipolar digital signal D [17]. Fig. 2.11 shows the function of the receiver.

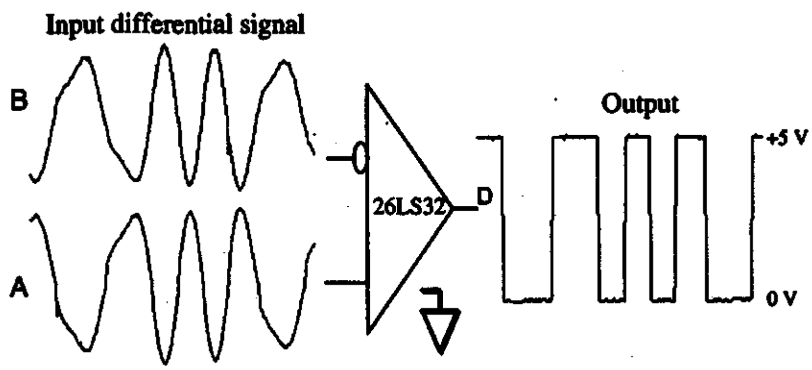


Fig. 2.11 The function of the LocalTalk receiver

The receiver only senses the difference between input signals. If the amplitude of signal A is higher than that of signal B, the output is high voltage, otherwise the output is low level voltage. The balanced configuration gives Apple LocalTalk some immunity to ground noise currents and common mode electromagnetic interference. Actual measurement recorded in Fig. 2.12 shows how the balanced receiver reduces the interference of the common mode noise. In Fig. 2.12, Signal A and signal B are the input signals of the receiver. Signal C shows the differences between inputs A and B. There is almost no common mode noise appearing in the signal C.

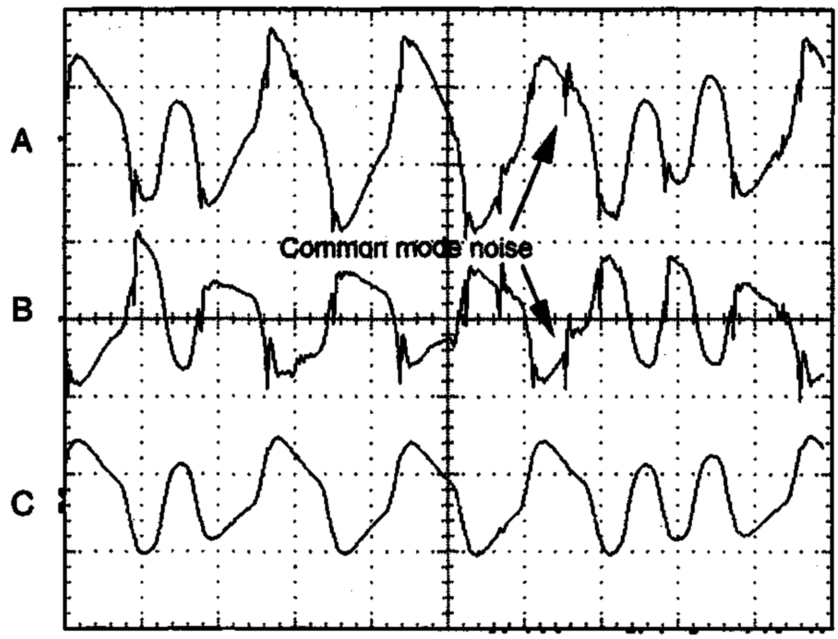


Fig. 2.12 Common mode noise rejection

Because of this advantage, the LocalTalk, modified from the RS-422 standard, can transmit data over longer distances with higher speed than the popular RS-232 interface standard which modulates a signal with respect to a common ground.

2.3.3 Serial Communication Controller (SCC)

The Zilog 8530 integrated circuit is used as the LocalTalk serial communication controller. It performs physical layer functions and some data link layer functions as well. Bit encoding/decoding, synchronization and carrier sense are carried out by this controller.

The baseband bit encoding/decoding used in LocalTalk is biphase space (or FM-0). The waveform is illustrated in Fig. 2.13. Note that each bit cell has a transition at its end. The receiver uses this as synchronization information. There is an extra transition at the middle of 0 bits. The Controller uses this mid-bit transition to decide whether the bit it receives is a 1 (mark character) or a 0 (space character).

Fig. 2.13 Biphase space signal

The SCC contains a Digital Phase Locked Loop (DPLL) to recover clock information from the FM-0 data stream. The DPLL is driven by a clock which is about 16 times the FM-0 data rate [18]. The DPLL uses this clock, along with the data stream, to construct a synchronized clock which is used to recover the received data.

The last function of physical layer is carrier sense. It is carried out by the SCC. The SCC can detect whether there is a device on the net that is transmitting data, or whether the clock of the data stream is missing. It reports such information to the Link Access layer for further processing.

2.3.4 Spectrum of the Apple LocalTalk signal

If an infinite length sequence of random binary data is assumed, after biphase encoding, it has the normalized power spectral density as in Eq. 2.1 [19].

$$P(f) = A^2T\left[\frac{\sin\left(\frac{\pi fT}{2}\right)}{\left(\frac{\pi fT}{2}\right)}\right]^2 \sin^2\left(\frac{\pi fT}{2}\right) \qquad \text{Eq. 2.1}$$

where: $P(f)$ is power spectral density (W/Hz).
A is amplitude of signal (V).
T is bit time (s).
f is frequency (Hz). Range is $-\infty < f < \infty$

For the 230.4 kbits/s LocalTalk data, T is 4.34 μs. If the amplitude A is 10 V, the normalized power spectral density of the signal is illustrated in Fig. 2.14

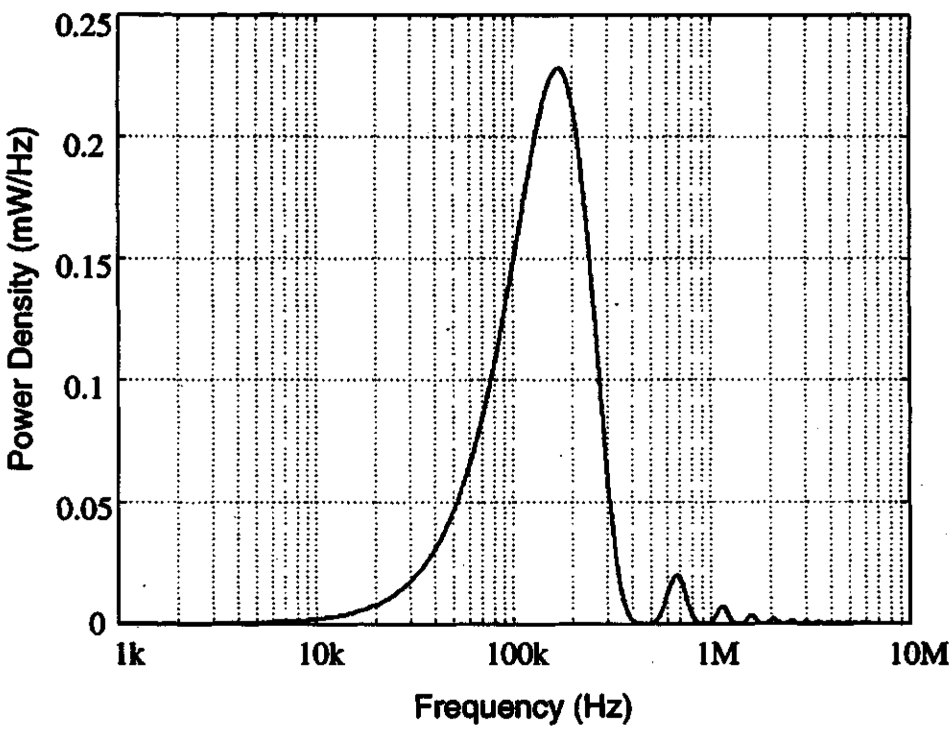


Fig. 2.14 Power spectral density (showing only positive frequency)

The power spectral density of infinite random binary sequence with biphase space encoding has some interesting characteristics.

The power in a specific frequency range can be calculated by Eq. 2.2.

$$E = 2\int_{f_1}^{f_2} P(f)df$$

$$= 2\int_{f_1}^{f_2} A^2 T \left[\frac{\sin\left(\frac{\pi f T}{2}\right)}{\left(\frac{\pi f T}{2}\right)} \right]^2 \sin^2\left(\frac{\pi f T}{2}\right) df$$

Eq. 2.2

Power calculated in different bands is listed in Table 2.1.

f_1 (kHz)	f_2 (kHz)	Normalized power (W)	Percent of total (ie. 100W)
0	4	0.00086	0.00086%
4	30	0.3562	0.3562%
30	100	10.8253	10.8253%
100	150	18.9133	18.9133%
150	200	22.4102	22.4102%
200	250	18.2336	18.2336%
250	300	10.3084	10.3084%
300	350	3.7681	3.7681%
350	400	0.7179	0.7179%
400	460.8	0.0367	0.0367%
460.8	691.2	4.0065	4.0065%
691.2	Infinity	10.4229	10.4229%

Table 2.1 The power distribution of the LocalTalk data

The power distributed in the frequency band from dc to 4 kHz, including the voice band (300 ~ 3300 Hz), is about 0.86 mW (or -0.66 dBm), which is only 0.00086% of the total power. The power is so small that a high pass filter can be used to eliminate it and this will not cause any significant distortion to the data signal.

The power of the signal in the frequency band from 30 kHz to 350 kHz has up to 89.22% of the total power. The first null in the frequency spectrum occurs at 460.8 kHz and the power beyond 460.8 kHz is only about 14.43% of the total power. Thus, the main power of the signal is concentrated between 30 kHz to 350 kHz and the components in this frequency band should have as little distortion as possible.

2.4 LocalTalk Link Access Protocol (LLAP)

The main tasks of LocalTalk Link Access Protocol are to provide link access management, data framing, data frame transmission/reception and node addressing. These tasks are conducted by the SCC and associated software. The objective of this protocol is to allow all devices on the LocalTalk network to share the same transmission media and to turn the noisy physical channel to an errorless transmission line for the network layer. The devices on the network which can perform this protocol are called nodes. All nodes should be within a stand alone network (see Fig. 2.2).

2.4.1 Link Access Management

All LocalTalk devices must be controlled by an access protocol to share the common physical channel, otherwise the data transmission can not be reliable and efficient. The LocalTalk Link Access Protocol uses Carrier Sense Multiple Access with

Collision Avoidance(CSMA/CA). The hardware of LocalTalk can sense the carrier when other nodes are transmitting, but it can't detect a collision. Collision avoidance is realized by requiring the transmitter to wait for a randomly generated amount of time after the bus has been sensed idle. Only after that random time can the transmitter send its Request-to-Send control frame.

2.4.2 Data Framing

Information that is transmitted between data link layers is organized as frames. The Apple LocalTalk link access protocol frame is shown in Fig. 2.15.

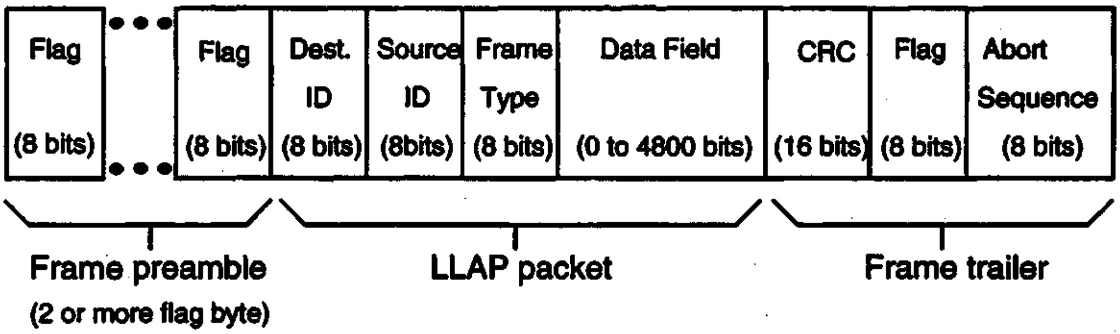


Fig. 2.15 LLAP frame

LocalTalk link access protocol uses bit oriented framing. Each frame starts with an 8-bit sequence (01111110) called a flag byte and each frame ends with the same flag byte. The receiver uses these flag bytes to define the data packet boundaries. Therefore the flag bit pattern is definitely not allowed inside a data stream. In order to avoid the flag byte sequence occurring inside the data stream which has arbitrary bit patterns, bit stuffing is necessary. When transmitting data, if 5 consecutive 1s are counted by the sender's link access protocol, a 0 is automatically stuffed into the outgoing stream. The receiver's link access protocol performs the process vise versa. When 5 consecutive 1s followed by one 0 is counted, the receiver deletes the 0 from the received data stream automatically.

Following the starting flag bytes, the frame has a destination ID byte and a source ID byte which are the 8 bit addresses of the receiver and sender. The frame type byte classifies the frames as control frames or data frames. The control frame doesn't have a data field. There are two pairs of control frames used in LLAP. One pair is called enquiry control (ENQ) and acknowledgment control (ACK) frame. They are used in addressing (detailed in Section 2.4.4). The other control pairs are called Request-to-Send (RTS) and Clear-to-Send (CTS) frames. They are used to control data packet transmission (see 2.4.3).

The 16 CRC-CCITT frame check sequence is applied to the LLAP packet. The abort sequence indicates the end of the frame. All framing operations (adding/deleting flag bytes, bit stuffing and CRC calculation/check) are accomplished by the serial communication controller (SCC).

A synchronization pulse is transmitted before a LLAP RTS frame. Fig. 2.16 shows the sync pulse.

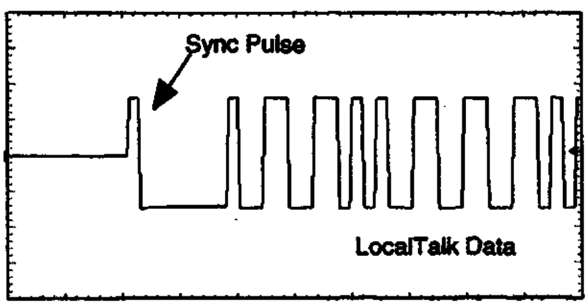


Fig. 2.16 Start of a LocalTalk RTS frame

All receivers on the network can take the transition of the synchronization pulse as a clock, but shortly they will lose this clock due to the following idle period of sufficient length. The missing clock detected by the receivers allows transmitters to synchronize their access to the line and know right away that another transmission is about to take place. The LLAP uses the synchronization pulse to help LocalTalk hardware detect the channel status and to reduce the probability of collision.

2.4.3 Frame Transmission/Reception

Directed and broadcast transmission dialogs are two types of frame transmission in LLAP. When two specific nodes in the network exchange information directly with each other, it is called a directed transmission dialog, while a node sending its data frame to all nodes on the network is called a broadcast dialog. There is only one dialog allowed on the channel at one time. Between two dialogs, there must be at least 400 μs gap (called inter-dialog gap or IDG). Different frames in one dialog must be separated by less than 200 μs (called interframe gap or IFG)

The implementation of the CSMA/CA protocol in a directed transmission dialog is illustrated in Fig. 2.17. The implementation of a broadcast transmission dialog is similar to that of a directed transmission dialog.

The operation of a receiver in the destination node is relatively simple. A node on the network can receive broadcast packets or a frame whose destination address matches the node's ID. Any bad frames which are caused by invalid frame type, synchronization error or an incorrect CRC will be rejected by the receiving node.

According to the directed frame transmission dialog, the maximum valid range for performing LLAP can be calculated. After transmitting the RTS frame, a sender must receive the CTS frame within the maximum inter-frame gap (T_{IFG}). The IFG (inter-frame gap) starts from the end of the abort sequence of the previous frame's trailer to the first bit of the current frame. Therefore, Eq. 2.3 must be satisfied.

$$T_{IFG} \geq T_{RTD} + T_P \qquad \text{Eq. 2.3}$$

where: $T_{IFG} = 200\ \mu s$
T_{RTD} is the round trip delay between the sender and the receiver node.
T_P is the time needed for receiver to answer the received RTS frame.

From Eq. 2.3

$$T_{RTD} \leq T_{IFG} - T_P \qquad \text{Eq. 2.4}$$

According to measurements in our laboratory, the longest time needed for the receiver to answer a RTS frame (T_P) is about 104 µs. Substituting T_P with 104 µs and T_{IFG} with 200 µs, T_{RTD} can be calculated from Eq. 2.4 as

$$T_{RTD} \leq 96\ \mu s \qquad \text{Eq. 2.5}$$

The speed of a signal that is transmitted in the twisted pair cable is about 2/3 of light speed (c) [20]. The maximum T_{RTD} is 96 µs, thus the maximum valid range (L_{max}) that the LLAP can perform correctly is given by

$$\begin{aligned} L_{max} &= \frac{\frac{2}{3} \times c \times \max(T_{RTD})}{2} \\ &\approx 9.6\ km \end{aligned}$$

Eq. 2.6

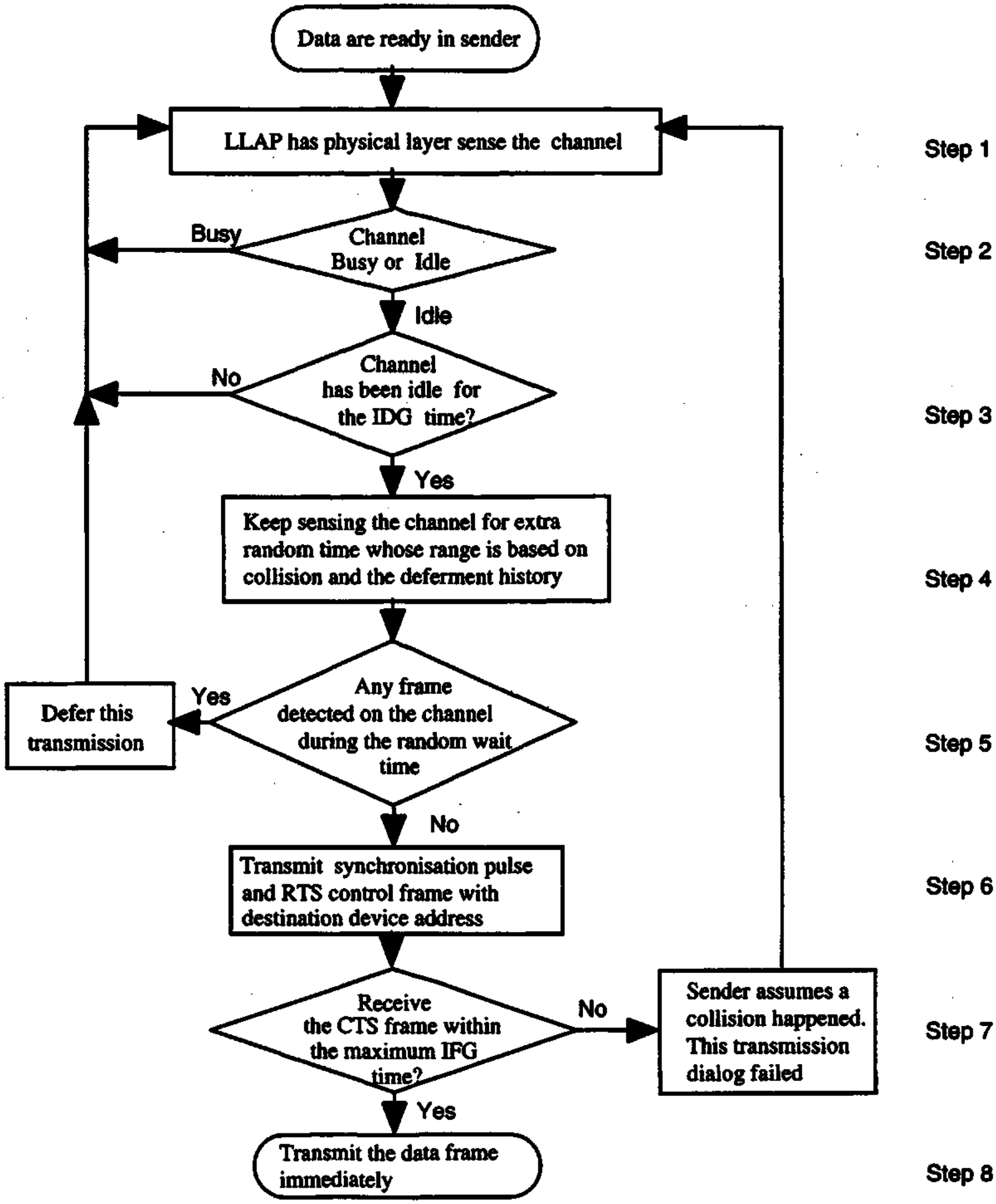


Fig. 2.17 Directed data transmission dialog

2.4.4 Node Addressing

LocalTalk uses a dynamic node address assignment scheme. Each node in the network has an 8-bit address that ranges from 1 to 254. When a new node is connected to the network, it randomly picks an initial address. In order to verify that this address is not used by other nodes, the new node sends an enquiry control frame (ENQ) to the initial address. If an acknowledgment control frame (ACK) is received, the new node knows that the initial address is already used and it has to pick another address and verify it again. If no response is received, this node can use this initial address as its address.

2.5 Network Layer and Other Upper Layers

The network layer in AppleTalk uses Datagram Delivery Protocol (DDP). While the LLAP protocol exchanges frames over a stand-alone network, DDP provides datagrams delivery over an AppleTalk internet (see Fig. 2.3 and Fig. 2.4). Network addressing, packets circulating and packets routing algorithm are considered by DDP.

Transport layer in AppleTalk can change the unreliable DDP datagram service to a loss-free delivery [12]. This function gives upper layer a reliable data service. It also provides routing table maintenance protocol for DDP routing algorithm, echo protocol for network test and name binding protocol for easier access to the network.

Session layer is built on transport layer. It provides session establishment, maintenance and cancellation. This layer also provides printer access protocol for work stations to use printer service, Data Stream protocol for a reliable duplex data delivery, and zone information protocol for network access management.

The presentation layer is built on the session layer. The AppleTalk presentation layer mainly provides AppleTalk filing protocol for remote file access.

Application layer usually refers to the application programs that use the network services supported by AppleTalk.

Chapter 3: Transmission Line Theory

A transmission line can carry electrical signals over a certain distance. When the length of the transmission line is longer than 1/8 wave length, the transmission of the signal should be analyzed using transmission line theory [21]. Data transmission at 230 kbits/s on the telephone subscriber line must be analyzed with transmission line theory. The speed and distance of data communication are limited by factors such as signal attenuation and signal reflections.

3.1 Transmission Line Model

In practical cases, two transmission lines usually run in parallel and are separated with a constant space, therefore a relatively consistent capacitance exists between the two lines. There is a small amount of leakage in the insulation between the two wires and causes conductance in every unit length of the line. The conductors of the line have series resistance and inductance distributed on the line. Since the transmission line has distributed electrical elements, it can not be simply modeled as a single circuit.

The transmission line model is shown in Fig. 3.1, where dx is an infinitely short length of the line and x is measured from the load toward the sending end of the small section dx. For simplicity, the inductance and resistance in the two wires are combined and the model is presented with a ground reference. The distributed elements per unit length are l (inductance), g (conductance) r (resistance) and c (capacitance). These elements are fundamental electrical characteristics of the transmission line, so they are also called primary constants. Voltage **V** and current **I** are functions of x. Both **V** and **I** are vectors with magnitude and phase angle.

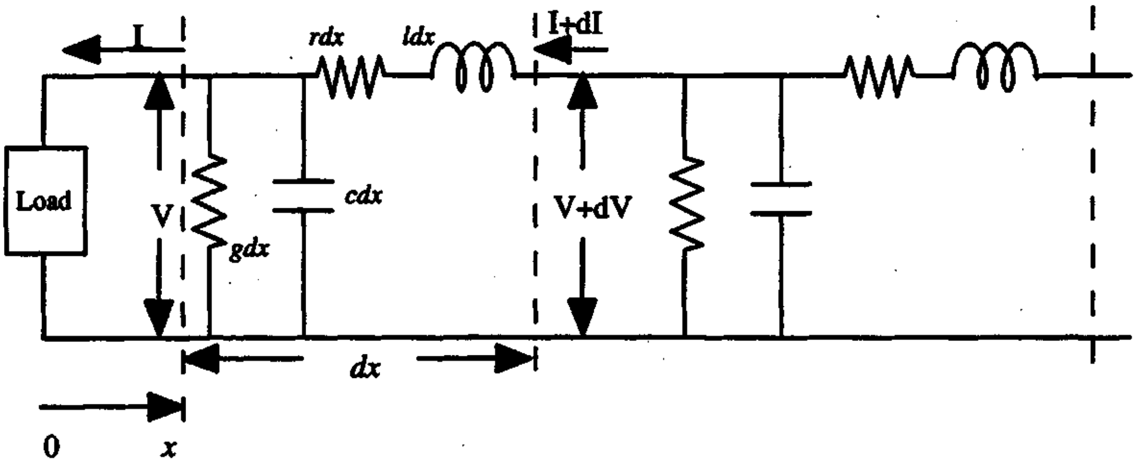


Fig. 3.1 Transmission line model

3.2 Circuit Analysis of the Model

The distributed character of the line causes the current and voltage to change continuously along the transmission line. Current at one end of a short section of line differs from current at the other end because of the leakage current through the distributed conductance and the shunted current through the capacitance from wire to wire. Voltage also varies from point to point because of series inductance and resistance.

Because the line model has both series and shunt parameters distributed along the circuit, it is necessary to analyze the transmission line model starting from any infinitesimal length of the line (dx). The voltage drop on dx is

$$d\mathbf{V} = (rdx + j\omega ldx)\mathbf{I}$$

or

$$\frac{d\mathbf{V}}{dx} = (r + jwl)\mathbf{I} \qquad \text{Eq. 3.1}$$

where ω is the radian frequency.

The current change in section dx is

$$d\mathbf{I} = (gdx + j\omega cdx)\mathbf{V}$$

or

$$\frac{d\mathbf{I}}{dx} = (g + j\omega c)\mathbf{V} \qquad \text{Eq. 3.2}$$

The voltage and current at point x can be obtained by solving Eq. 3.1 and Eq. 3.2:

$$\mathbf{V} = \mathbf{V}_1 e^{x\sqrt{zy}} + \mathbf{V}_2 e^{-x\sqrt{zy}} \qquad \text{Eq. 3.3}$$

$$\mathbf{I} = \mathbf{I}_1 e^{x\sqrt{zy}} + \mathbf{I}_2 e^{-x\sqrt{zy}} \qquad \text{Eq. 3.4}$$

where $z = r + j\omega l$, $y = g + j\omega c$ and x is the line length starting from the load.

The physical meaning of Eq. 3.3 and Eq. 3.4 is very important in the transmission line theory. Eq. 3.3 can be explained as the voltage on a transmission line is the sum of two traveling waves, one is the incident wave which travels toward the load and the other is the reflected wave which is sent back from the load. Eq. 3.4 shows that the current is also the sum of an incident wave and a reflected wave. $\mathbf{V}_1$ and $\mathbf{V}_2$ are incident and reflected voltages at the load end ($x = 0$). $\mathbf{I}_1$ and $\mathbf{I}_2$ are the corresponding currents of $\mathbf{V}_1$ and $\mathbf{V}_2$. Some important transmission line parameters can be obtained from Eq. 3.3 and Eq. 3.4

Characteristic Impedance:

The relation between V_1 and V_2 and their corresponding currents I_1 and I_2 can be obtained by substituting Eq. 3.3 and Eq. 3.4 into Eq. 3.1 and Eq. 3.2 [22]. The results are

$$V_1 = \sqrt{\frac{r + j\omega l}{g + j\omega c}} I_1 \qquad \text{Eq. 3.5}$$

and

$$V_2 = -\sqrt{\frac{r + j\omega l}{g + j\omega c}} I_2 \qquad \text{Eq. 3.6}$$

Eq. 3.5 and Eq. 3.6 show that the incident and reflected voltages and currents are related by $\sqrt{\frac{r + j\omega l}{g + j\omega c}}$, called characteristic impedance. Usually, characteristic impedance can be written as

$$Z_0 = \sqrt{\frac{z}{y}} \qquad \text{Eq. 3.7}$$

where $z = r + j\omega l$ and $y = g + j\omega c$

Eq. 3.5 and Eq. 3.6 can be rewritten as

$$V_1 = Z_0 I_1 \qquad \text{Eq. 3.8}$$

$$V_2 = -Z_0 I_2 \qquad \text{Eq. 3.9}$$

Each type of transmission line has its own characteristic impedance. It is based on the distributed elements of the transmission line and not on the length of the line.

Propagation Constant

The propagation constant γ is defined as Eq. 3.10,

$$\gamma = \sqrt{zy} \qquad \text{Eq. 3.10}$$

Since both z and y are complex numbers, the propagation constant is a complex number as well. It has a real part α and an imaginary part $j\beta$,

$$\gamma = \alpha + j\beta \qquad \text{Eq. 3.11}$$

The real part is called the attenuation constant, which quantifies the voltage or current attenuation on the transmission line. The imaginary part is called the phase constant, which is responsible for phase shift when the signal is measured at points along the line. The delay at certain frequency can be calculated by

$$Delay = \frac{\beta}{\omega} \qquad \text{Eq. 3.12}$$

where β is the phase shift, ω is the radian frequency.

Because delay depends on the frequency, different frequency waves have different transmission delay and therefore the delay has a nonlinear relation with frequency.

Since characteristic impedance (Zo) and propagation constant (γ) are derived from the primary constants and are also used to describe the transmission line characteristics, they are called secondary parameters. The primary constants and the secondary parameters are temperature dependent. The conductor's resistance changes with temperature, so the attenuation is higher when temperature is higher.

The Reflection Factor

The voltage or current at every point of the line is the sum of the incident and reflected voltages or currents. The reflection level of the reflected signal can be measured as the voltage ratio between reflected signal and the incident signal at the receiving end ($x = 0$). This ratio is called the reflection factor.

According to definition, the reflection factor can be written as

$$\rho = \frac{V_2}{V_1} \qquad \text{Eq. 3.13}$$

where V_1 is the incident voltage at the receiving end and V_2 is the reflected voltage.

The voltage (V_r) and current (I_r) at the receiving end can be obtained by substituting $x = 0$ to Eq. 3.3 and Eq. 3.4.

$$V_r = V_1 + V_2 \qquad \text{Eq. 3.14}$$

$$I_r = I_1 + I_2 \qquad \text{Eq. 3.15}$$

The impedance at the receiving end is

$$Z_r = \frac{V_r}{I_r} \qquad \text{Eq. 3.16}$$

To combine Eq. 3.16, Eq. 3.15, Eq. 3.14, Eq. 3.9 and Eq. 3.8 together, the relation between V_1 and V_2 can be obtained as

$$\frac{V_2}{V_1} = \frac{Z_r - Z_0}{Z_r + Z_0} \qquad \text{Eq. 3.17}$$

so the reflection factor can also be calculated as

$$\rho = \frac{Z_r - Z_0}{Z_r + Z_0} \qquad \text{Eq. 3.18}$$

3.3 Termination of the Transmission Line

When the transmission line is terminated by a load whose impedance Z_r is equal to the characteristic impedance Z_0, the reflection factor ρ is 0 (see Eq. 3.18). This means there is no reflection at the load. This is called impedance matching. If the transmission line is terminated by any load whose impedance is not equal to Z_0 in phase and magnitude, a reflection will appear.

One extreme impedance mismatch case is when the transmission line is an open-circuited line which has no load at the end. The bridged tap, which will be discussed in Chapter 6, is similar to this case so it is necessary to know how waves travel in the open-circuited line.

Because the load of the open circuited line is infinitely large ($Z_r = \infty$), the limit of the reflection factor is 1. This means that the total incident wave is reflected back to the source. The voltages of waves at different points on an open-circuited line are shown in Fig. 3.2.

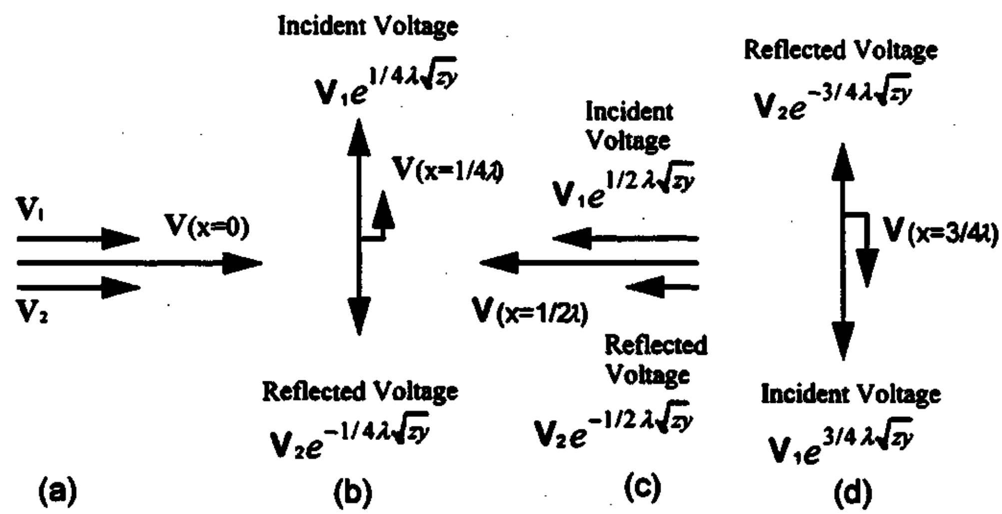


Fig. 3.2 Voltages at different points on a line

In Fig. 3.2a, V_1 and V_2 are the incident and the reflected voltage at the load end (x = 0). At this point, V_1 and V_2 are equal in phase and magnitude. Voltage $V_{(x=0)}$ is the sum of V_1 and V_2.

Fig. 3.2b shows the voltages at a point of 1/4 wavelength from the load end. The reflected voltage at this point lags 90 degrees with respect to the reflected voltage at the load end. The magnitude of the reflected voltage at this point is also smaller than the reflected voltage at the load point because of the attenuation of the transmission line. The incident voltage at this point leads the incident voltage at the load point by 90 degrees and has larger magnitude. Therefore the two vectors are opposite to each other and result in a small sum voltage $V_{(x=1/4\lambda)}$.

At 1/2 wavelength from the load point, the incident voltage and the reflected voltage are in the same phase again but have 180 degrees difference to the voltages at the load end (shown in Fig. 3.2c).

Fig. 3.2d shows the vectors of the incident voltage and the reflected voltage at 3/4 wavelength point. The phases of the incident voltage and the reflected voltage have 180 degrees difference. Again the sum of the two vectors is small.

According to Fig. 3.2, the voltage attenuation can be high if the point is at n/4 wavelength, where n is an odd number.

In low speed data transmission, the wavelength of the data is long, the reflected signals have to travel longer distance and lose more power to get the 1/4 wavelength point. Therefore, a small amount of reflected signal is allowed and it is not necessary to terminate the transmission line by a load whose impedance exactly equals Z_0. Practically, impedance matching for the low speed data is usually realized by using only a matched resistor at the load, such as the termination used in LocalTalk network which is discussed in Chapter 2. However, the wavelength of the high speed data is very short, the reflected signals lose little power to get the 1/4 wavelength point and a small amount of reflected signals can seriously interfere with the incident signals. As a result, the impedance matching in phase is as important as in magnitude.

3.4 Transmission Cables in Telephone Networks

Transmission cables play very important roles in telecommunication systems. Local telephone and cable TV networks are mainly connected by transmission cables. Because the signals are transmitted in an guided manner, these cables are also called a guided transmission medium [23]. According to the structure, transmission cables can be classified as open-wire lines, paired cables and coaxial cables.

3.4.1 Open Wire Lines

In the early years, telephone networks were connected by various forms of multiple line wires which were hung between telephone poles. These open wires were subject to cross talk and the effects of rain, frost, and ice on transmission [24]. Because of these disadvantages and the high cost of maintenance, most open wire lines have been replaced by coaxial cables and paired cables in modern telephone networks.

3.4.2 Coaxial Cable

The useable bandwidth of a coaxial cable is very high and up to 10,800 voice channels can be carried by a single coaxial unit [25]. Although the cost of the coaxial cable is high, the per-channel-mile cost is still low if it is used in a high traffic place. Thus, the coaxial cables are often used as toll cables in the telephone networks.

Because the inner conductor is surrounded by the cylindrical conductor, the coaxial unit has excellent immunity to cross talk and outside noise than the twisted pair when compared with twisted pair. The coaxial cable's attenuation is nearly proportional to the square root of frequency and the velocity of propagation and characteristic impedance vary slightly with frequency. Coaxial transmission line can be installed as a single coaxial unit or as a cable made up of 4 to 22 coaxial units (See Fig. 3.3). Each coaxial unit consists of a hollow cylindrical conductor surrounding a single wire conductor. The space between the cylindrical shell and the inner conductor is filled with a plastic insulator. A number of wires are packed in among the coaxial units and are used for maintenance support and control.

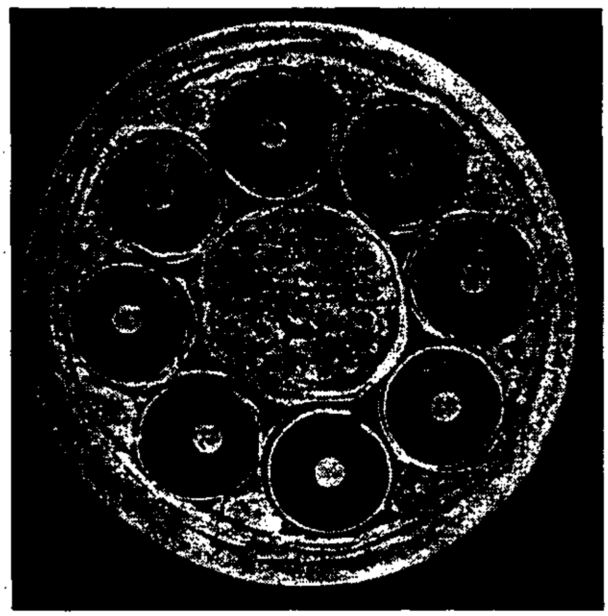


Fig. 3.3 Coaxial cable

(A copy from *Principles of Electricity applied to Telephone and Telegraph Work [24]*)

3.4.3 Paired Cables

Telephone subscriber station sets are connected with paired wire cables to local switching and transmission equipment. The average length of this subscriber transmission line is approximately 3 km. Cable conductors are insulated with polyethylene or pulp papers, and twisted in pairs. Groups of such pairs can be twisted into a rope-like unit. Several units are twisted together to form a cable core. This type of structure can provide equal interference exposure to each wire pair and minimize the electromagnetic interference from neighbor pairs, power lines and other sources. In order to prevent damage to cables from water penetration, rodents and corrosive elements in the soil, protective methods are necessary.

In North America, the size of the conductor in a paired cable is standardized as America Wire Gauge (AWG). Table 3.1 shows some most commonly used AWG conductor dimensions [21]. The primary constants of different size cables are not the same, so the propagation constants vary as shown in Fig. 3.4 and Fig. 3.5 [34].

AWG	19	22	24	26
Diameter (mm)	0.912	0.645	0.511	0.404

Table 3.1 AWG wire size

The attenuation of these cables is similar in the low frequency band but there are significant differences at high frequency. With frequency increasing, the attenuation of larger size cable (such as AWG 19) grows more slowly than that of smaller size cable (such as AWG 26). Fig. 3.5 shows that the delay curves of different size cables are almost the same at higher frequency, but the larger size cable has smaller delay in the low frequency band. These two figures indicate that the larger size cable has a wider bandwidth and is suitable to high speed data transmission and carrier systems. In the telephone network, AWG 26 PIC (polyethylene insulated cable) is mostly used in a subscriber loop whose main purpose is to transmit voice. AWG 24 and AWG 22 cables are used in 1.544 Mbits/s T1 trunks or other types of toll trunks [21].

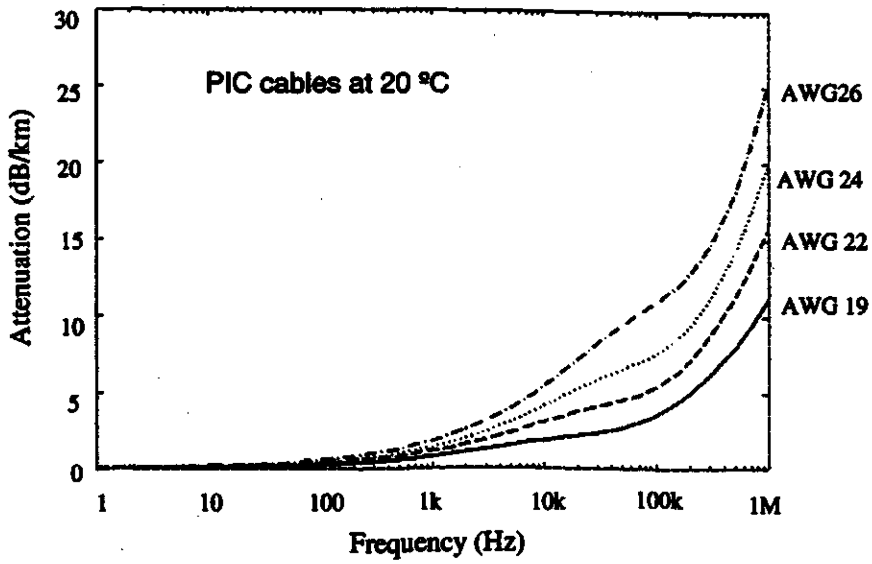


Fig. 3.4 Attenuation of different size cables vs. frequency

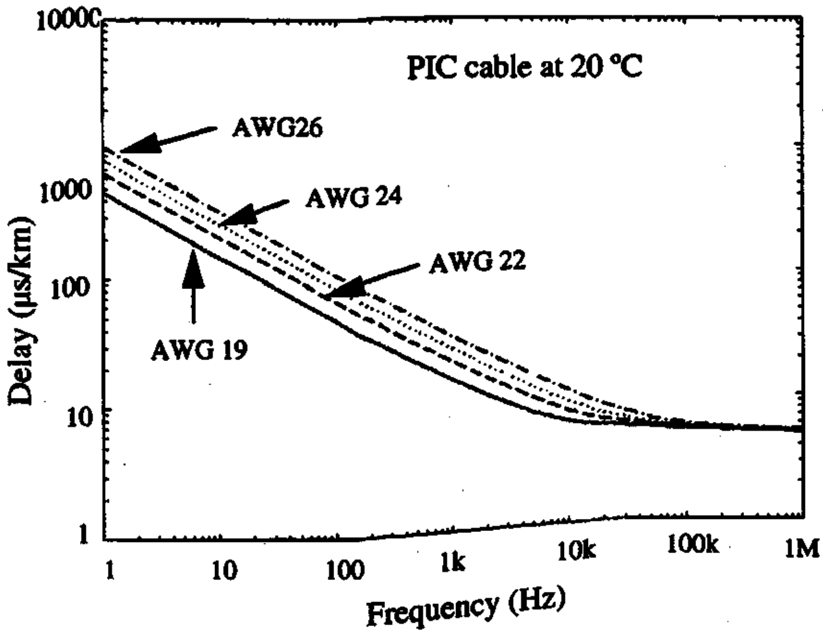


Fig. 3.5 Delay of different size cables vs. frequency

As mentioned before, the propagation constants change with temperatures. The propagation constants of AWG 26 PIC cable at different temperatures are illustrated in Fig. 3.6 and Fig. 3.7 [34].

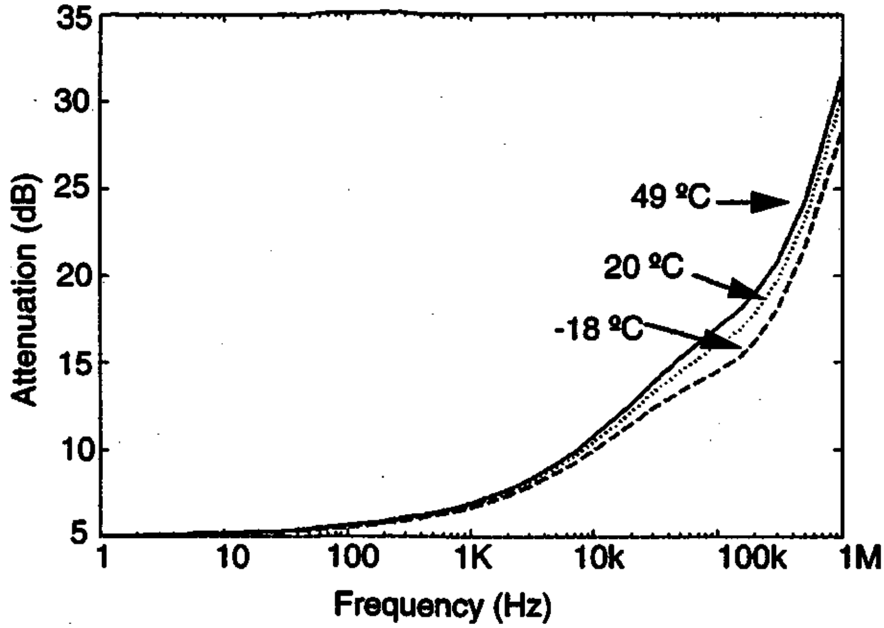


Fig. 3.6 Attenuation of AWG 26 cable at different temperature

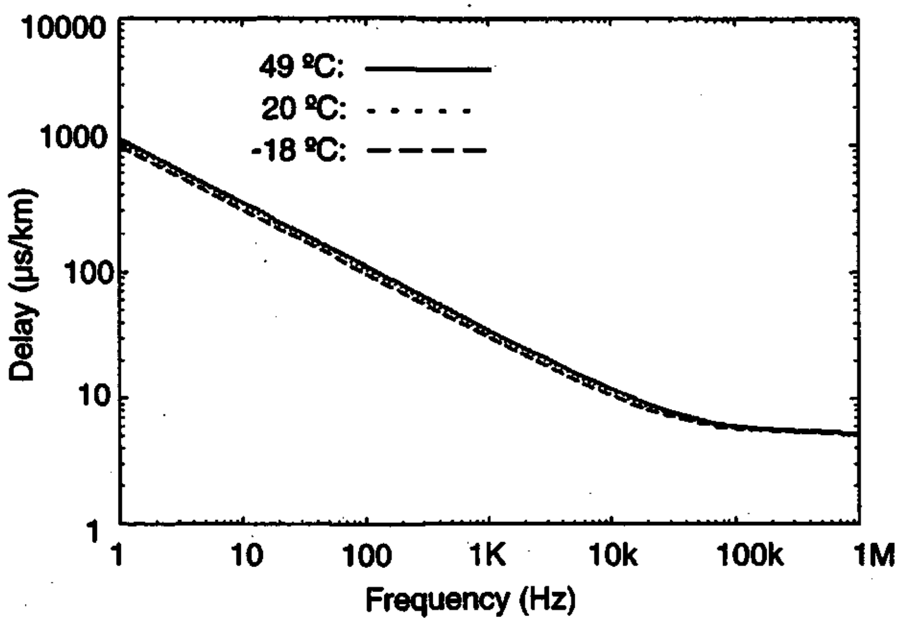


Fig. 3.7 Delay of AWG 26 cable at different temperature

Fig. 3.7 and Fig. 3.8 show that the effects of the temperature are not serious. The delay factor is almost unchanged and the attenuation only has a maximum difference of 3 dB from -18 °C to 49 °C.

The characteristic impedance of the cable also changes with frequency. For example, Fig. 3.8 and Fig. 3.9 shows the characteristic impedance of the AWG 26 cable at 20 °C.

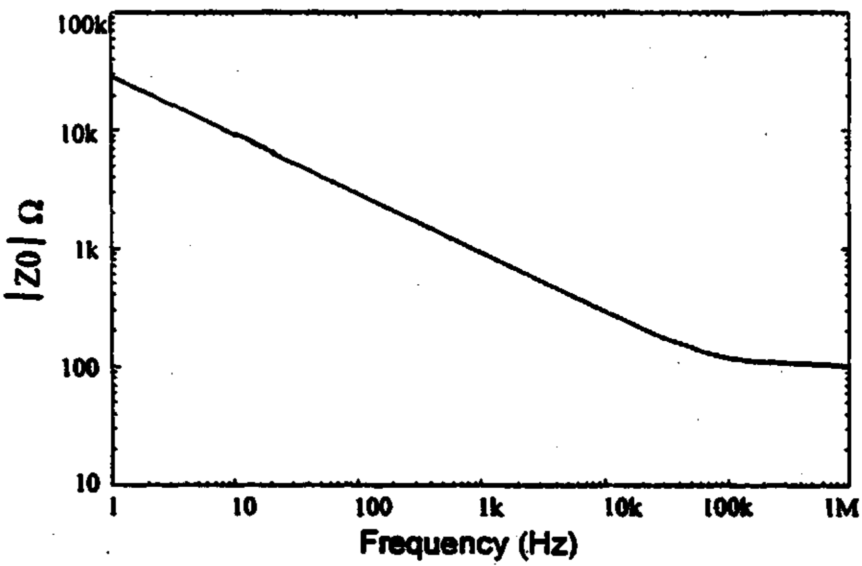


Fig. 3.8 Characteristic impedance of AWG 26 cable vs frequency

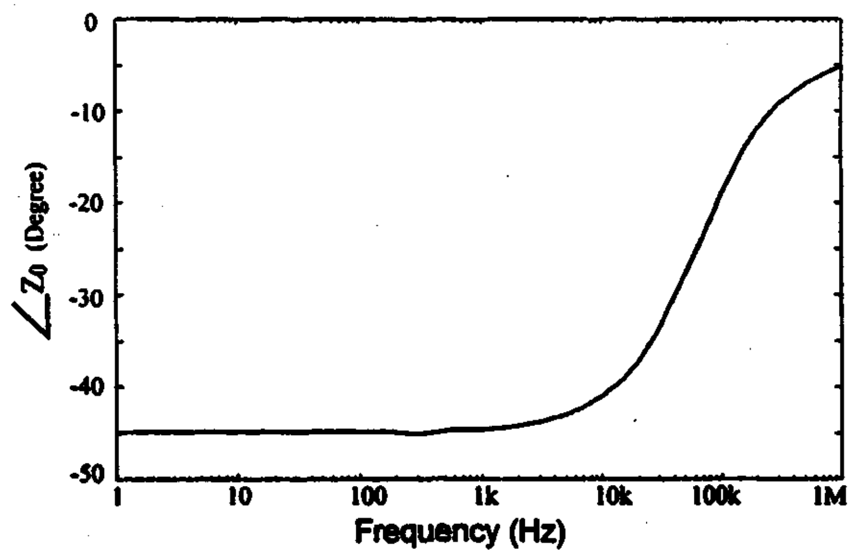


Fig. 3.9 Angle of the characteristic impedance vs frequency

The characteristic impedance of the cable is high in the low frequency band and changes to about 100 Ω with increasing frequency. The angle of the characteristic impedance has a sharp curve from 10 kHz to 1 MHz, so it is not easy to completely match the cable's characteristic impedance in magnitude and angle over a large frequency band.

In this project, data are transmitted in an AWG 26 cable. This type of cable will cause delay distortion (or phase distortion) and attenuation distortion in the data signal. The distortion caused by the different attenuation between different frequencies is called attenuation distortion. As shown in Fig. 3.4, the AWG 26 cable has the largest attenuation distortion of the cable pair family. Delay distortion is caused by the lower frequencies having longer delays than the higher frequencies (See Fig. 3.7). Since the delay will create incorrect relative phase, it is also called phase distortion. The difference in delay between different frequency components increases with transmission line length. As shown in Fig. 3.10A, Curve 1A has two harmonics, a and b. If the delays of harmonics a and b are not the same, a distorted Curve 1B occurs as shown in Fig. 3.10B.

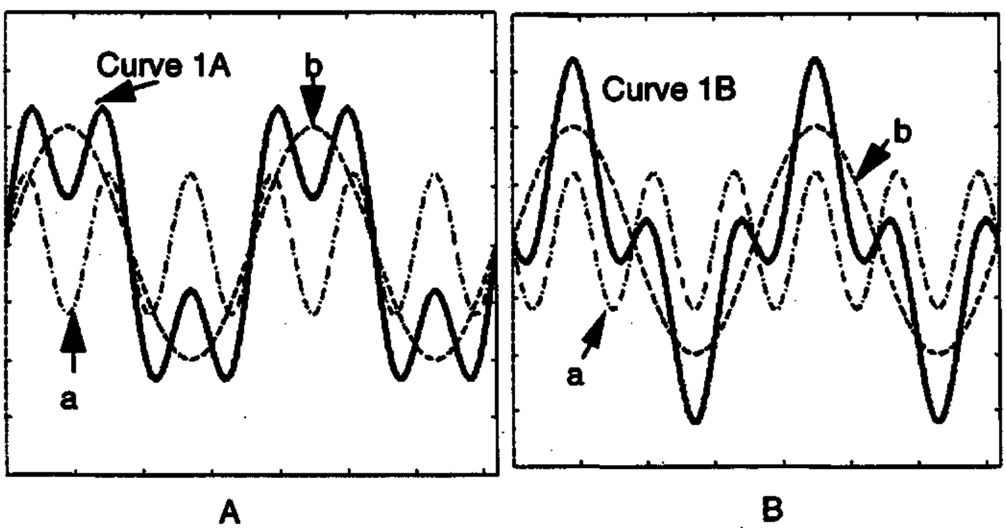


Fig. 3.10 Illustration of delay distortion

Chapter 4 : Data Transmission

In passband transmission, a carrier can be used to modulate the data before the transmission. Alternatively the data can be sent directly without any modulation. and this is called baseband transmission. The passband transmission can move the spectrum of the data to a specific frequency range that is suitable for the requirement of the communication channel. However, the passband transmission needs a modulator in the transmitter and a demodulator in the receiver, therefore this system becomes relatively complex. The baseband transmission only needs a transmitter and a receiver, so it is simple and suitable for short distance communication. This project uses baseband transmission.

4.1 Baseband Transmission

According to different analyzing requirements, baseband transmission can be abstracted as coded baseband transmission or baseband pulse transmission.

4.1.1 Coded Baseband Transmission

A coded baseband transmission system can be block-diagrammed as in Fig. 4.1. The line encoder encodes binary data to one specific channel code whose bandwidth, spectrum, timing information, and error control are most suitable for the transmission. The encoded signals are transmitted to the channel by the line driver. The receiver and line decoder recover the binary data from the encoded signals at the end. The characteristics of the transmission channel and the receiver complexity determine which channel code is the best choice.

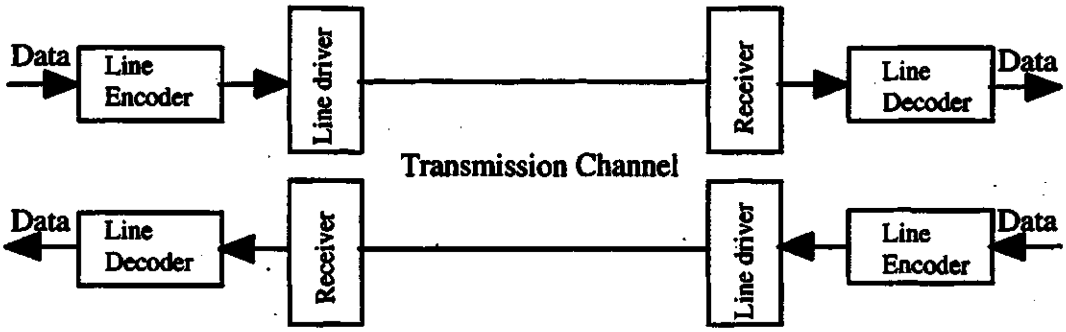


Fig. 4.1 Coded baseband transmission model

Several common used channel codes are illustrated in Fig. 4.2.

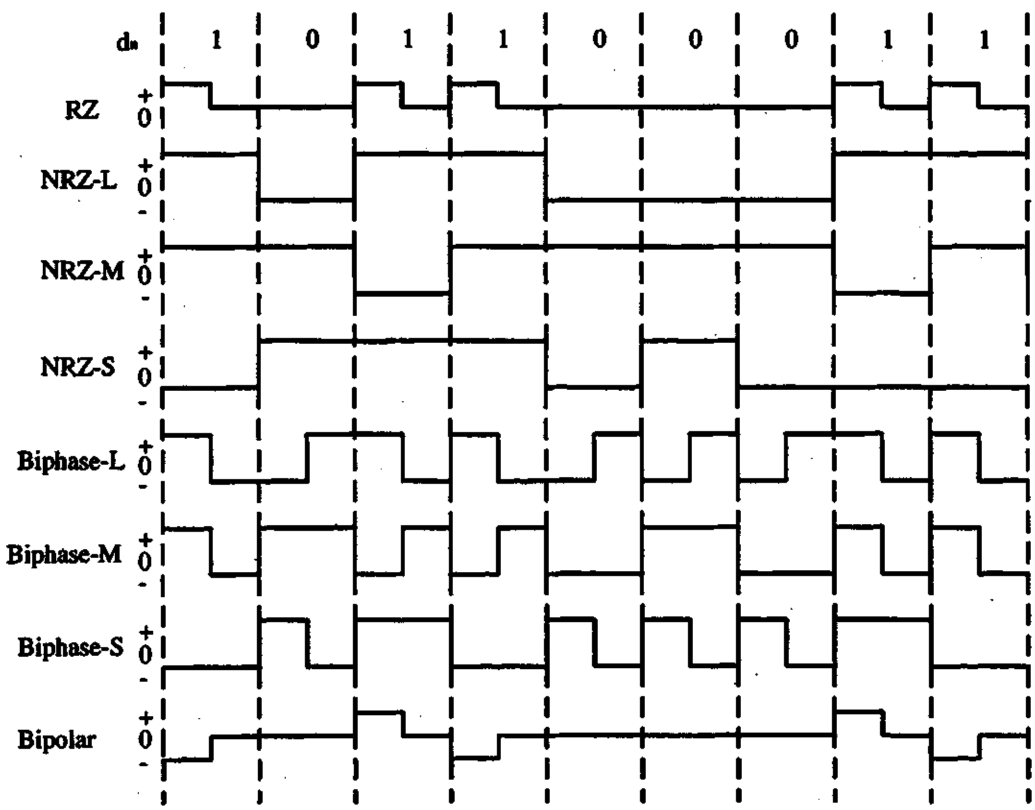


Fig. 4.2 Channel codes formats

Return to Zero (RZ) Code

A RZ code is an unipolar code. The binary 1 is represented by a positive pulse for half-bit period then returning to zero and a binary 0 is encoded to zero level. The advantage of this coding technique is relatively simple, but it doesn't have synchronization capability and has strong direct-current (dc) component. Because of the dc component in the signal, transmission equipment must be directly physically attached and can not be coupled via a transformer. The direct connection is vulnerable to interference due to the lack of electrical isolation.

Nonreturn to Zero (NRZ)

NRZ codes are polar codes. The binary 1 and 0 are represented by equal positive and negative voltages whose levels are constant during a bit interval. Because NRZ-L (NRZ level) is the simplest code to implement, it can be used to generate digital data by data processing terminals and can be changed to other types of code by the transmission system. The NRZ coding is very easy to be realized and has a narrow bandwidth, so is widely used in the data communication systems. However, the problems with the dc component and synchronization are still remained in NRZ signals. NRZ-S (NRZ space) and NRZ-M (NRZ mark) are differential NRZ codes.

In differential coding, the encoded differential data is generated by

$$e_n = d_n \text{ XOR } e_{n-1} \qquad \text{Eq. 4.1}$$

where e_n is the encoded sequence.
e_{n-1} has 1 bit delay to e_n.
dn is the input sequence.

The received data is decoded by

$$dn = e_n \text{ XOR } e_{n-1} \qquad \text{Eq. 4.2}$$

The advantage of differential codes may be illustrated by the differential encoding and decoding example shown in Fig. 4.3 [19].

Encoding								
Input sequence	d_n		1	0	1	1	0	
	e_{n-1}		0	1	1	0	1	
$e_n = d_n$ XOR e_{n-1}	e_n	0	1	1	0	1	1	Encoded sequence
Reference digit								
Decoding (with correct channel polarity)								
Received sequence	e_n	0	1	1	0	1	1	
	e_{n-1}		0	1	1	0	1	
$d_n = e_n$ XOR e_{n-1}	d_n		1	0	1	1	0	Decoded sequence
Decoding (with inverted channel polarity)								
Received sequence	e_n	1	0	0	1	0	0	
	e_{n-1}		1	0	0	1	0	
$d_n = e_n$ XOR e_{n-1}	d_n		1	0	1	1	0	Decoded sequence

Fig. 4.3 Example of differential coding

According to this example, an encoded sequence can be generated by inserting a reference digit to the first bit of the sequence and encoded by Eq. 4.2. For decoding operation, no matter whether the polarity of received sequence is correct or inverted, the received sequence can be always decoded correctly. This is a great advantage in the complex transmission system where the sense of the polarity of the signal may be easily lost.

Biphase

Biphase coding has several advantages over NRZ and RZ coding techniques. Because biphase coding requires at least one transition per bit interval, a receiver can use these predictable transitions during each bit time to synchronize the received data. Therefore, biphase codes are also known as self-clocking codes. The absence of an expected transition can be used to detect errors. Besides the advantages in synchronization and error detection, biphase codes have no dc components. This is desirable in many applications. The disadvantage of biphase codes are that they have wider bandwidths.

Biphase codes are widely used in network data transmission. For example, Ethernet uses Biphase-L (Manchester) codes and LocalTalk network uses Biphase-S (differential) codes.

Bipolar

Bipolar coding uses 3 encoded signal levels (+, 0, -) to represent binary data (0,1). Since successive 1's have alternative positive and negative polarities, bipolar provides some error-detection capability. Bipolar codes also have advantages of no dc component and having narrower bandwidth than biphase codes. In addition, Bipolar codes have some synchronization capability. T1-PCM transmission uses bipolar coding.

4.1.2 Baseband Pulse Transmission

A baseband pulse transmission system model is shown in Fig. 4.4.

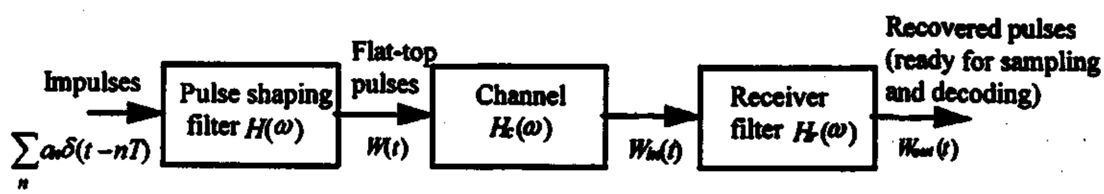


Fig. 4.4 Baseband pulse transmission system model

A series of impulses is input into the pulse shaping filter ($H(\omega)$) with the rate $1/T$. The pulse shaping filter modifies the impulses to flat-topped pulses ($W(t)$) and they are changed to $W_{in}(t)$ after the pulses go through the channel ($H_c(\omega)$,. $W_{out}(t)$ is the response of the receiver filter ($H_r(\omega)$) with the input of $W_{in}(t)$.

We may assume that the impulse response of the system is $h(t)$, but when only one impulse is input into the system, the output $W_{out}(t)$ can be treated as an impulse response of the system.

4.2 Intersymbol Interference

Based on Fig. 4.4, we assume $G(\omega)$ to be the spectrum of a pulse. The spectrum of the received pulse at the receiver is

$$W_{out}(\omega) = G(\omega) H_c(\omega) H_r(\omega) \qquad \text{Eq. 4.1}$$

If an undistorted pulse is wanted at the receiver, the following condition must be satisfied:

$$\begin{cases} |H_c(\omega) H_r(\omega)| = K & |\omega| \le \omega_c \\ |G(\omega)| = 0 & |\omega| > \omega_c \end{cases} \qquad \text{Eq. 4.2}$$

where ω_c is the channel cut off frequency.

However, the absolute bandwidth of a pulse is infinity. Therefore, if these pulses are passed to a communication system with limited bandwidth, they will smear out in the time domain and interfere with adjacent pulses. The interference between data pulses after being transmitted through a band limited channel is called the intersymbol interference (ISI).

Intersymbol interference is an important issue in data communication. It will worsen the noise immunity and cause decision errors at the receiver.

4.3 Equalization and Predistortion

In a practical transmission system, it is impossible to transmit data without distortion. The bandwidth of the transmission channel is always limited and the spectrum of the pulse extends to infinite frequency, so intersymbol interference is unavoidable. In order to achieve a less distorted signal, it is necessary to compensate for the magnitude and the phase characteristics of the transmission system. This kind of adjustment is called equalization if the adjustment is conducted to the received signal before the receiver, or predistortion if the adjustment is made before the signal is transmitted to the channel.

Usually the equalizer is placed in front of the receiver as shown in Fig. 4.5. When the equalizer is located at the receiver, the distortion of the line can be determined and automatically compensated by an adaptive equalizer. For this project, we are using a fixed equalizer that has lower cost and less complexity.

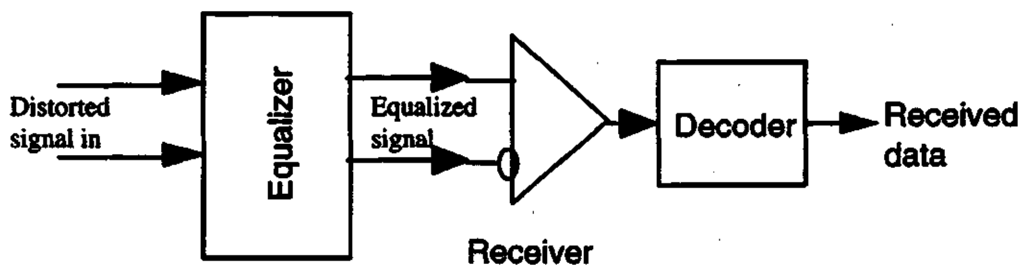


Fig. 4.5 The application of the equalizer

The predistorter can be placed after the line driver. The use of the predistorter can reduce the complexity of the equalizer at the receiver side and increase the performance of the system.

One specific predistorter that is used in this project will be described in Chapter 6. Two categories of equalizers are described in the following two sections.

4.3.1 Frequency Domain Equalization

The objective of frequency domain equalization is to compensate for the frequency dependent attenuation caused by the channel and to linearize the phase versus frequency response of the channel. The ideal attenuation and phase equalizations are shown in Fig. 4.6a and Fig. 4.6b. According to these two figures, we can see that the gain and delay responses of an ideal equalizer are inverse curves of the line loss and delay.

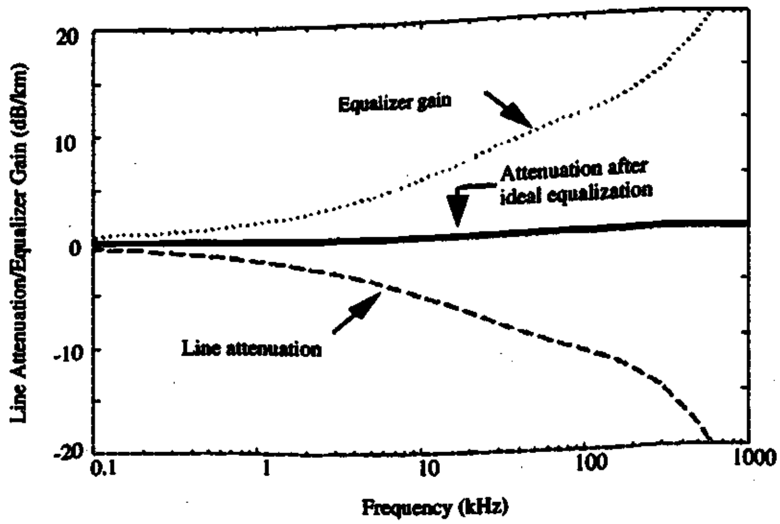


Fig. 4.6a Ideal equalization (Attenuation)

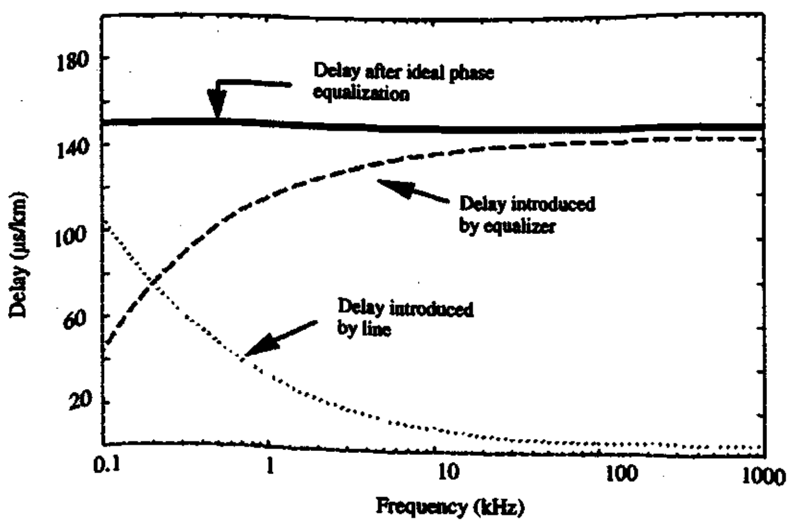


Fig. 4.6b Ideal equalization (Phase)

Unfortunately it is impossible to flatten the magnitude or linearize the phase over all frequencies. The equalizer can only compensate for the attenuation or linearize the phase in a certain frequency range. Usually, the equalization is conducted over the frequency range where the power of the signal is concentrated. For LocalTalk data, it has high power density from 30 kHz to 350 kHz. The frequency domain equalizer should be focused on this range.

Because frequency domain equalization must know the channel's frequency responses, so only those channels whose frequency responses are well known, such as twisted pair or coaxial cable, are suitable for frequency domain equalization. The compensation ability and precision of this kind of equalizer are not high, but frequency domain equalizers are efficient for use in the system where the bandwidth is not strictly limited and the signal distortion is not very serious. Although there are some restrictions in application, frequency domain equalizer has the advantages of simple design and low cost.

4.3.2 Time Domain Equalization

If the equalization is performed in the time domain, it is called time domain equalization. The key issue for time domain equalization is to minimize the intersymbol interference. There are a lot of algorithms available for time domain equalization [26]. The simplest one that can be used to illustrate how the equalizer works is the zero forcing algorithm. The objective of this algorithm is to let the system impulse response ($h(t)$) of the transmission system (see Fig. 4.4) satisfy the following conditions:

$$h(kT) = \begin{cases} C, & k = 0 \\ 0, & k \neq 0 \end{cases} \qquad \text{Eq. 4.3}$$

Where T is the symbol (sample) clocking period.

C is a non-zero constant and k is an integer ranged from $-\infty$ to $+\infty$.

If Eq. 4.3 is satisfied, a single flat-top pulse, generated from an impulse by the pulse shaping filter (see Fig. 4.4) going through the transmission system, would have a value C at $t = 0$ but would not cause interference ($h(t) = 0$) at other sampling times ($t = kT$, for $k \neq 0$) (see Fig. 4.7). Therefore intersymbol interference is eliminated at each sample moment.

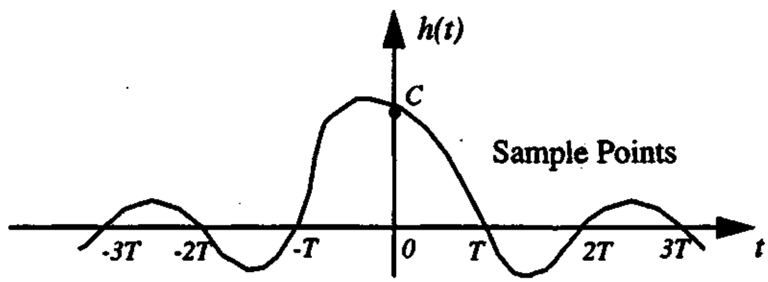


Fig. 4.7 Waveform with zero ISI

The zero forcing algorithm can be realized by a transversal filter which is often used for time domain equalization. As illustrated in Fig. 4.8, the transversal filter consists of a tapped delay line where the tap outputs are multiplied by a set of gains $\{C_n\}$. The resulting weighted tap outputs are summed to produce the filter output $W_{out}(t)$. The tap spacing is chosen to be equal to the symbol duration T or a fraction of the symbol duration.

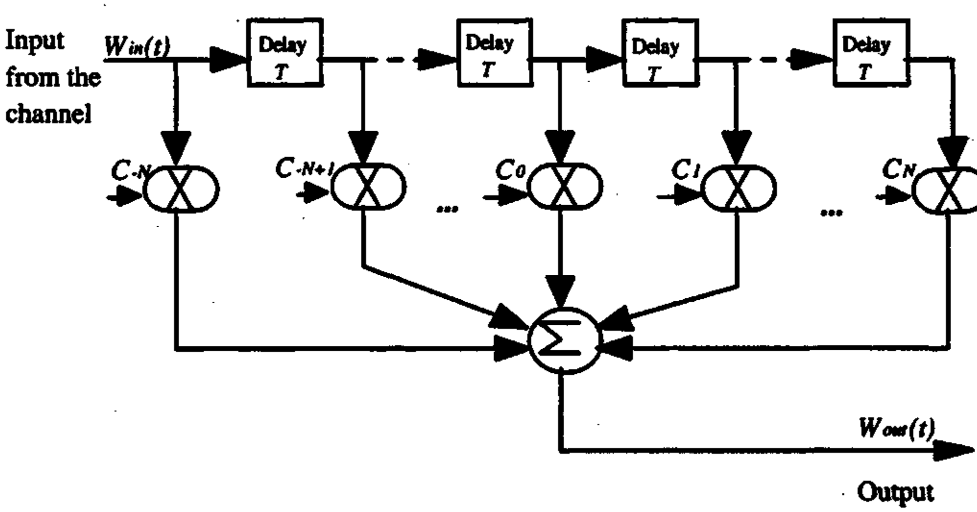


Fig. 4.8 Transversal filter

The key issue in designing the transversal filter is to find the set $\{C_n\}$ that can eliminates ISI. In order to get the $\{C_n\}$, we have to start from the output of the filter. The output of the filter is related to its input by

$$W_{out}(t) = C_{-N}W_{in}(t) + C_{-N+1}W_{in}(t-T) + \cdots + C_N W_{in}(t-2NT)$$

or

$$W_{out}(t) = \sum_{n=-N}^{N} C_n W_{in}(t-(n+N)T) \qquad \text{Eq. 4.4}$$

Evaluating the samples of $W_{out}(t)$ at the sampling times $t = kT$, we get

$$\begin{aligned} W_{out}(kT) &= \sum_{n=-N}^{N} C_n W_{in}(kT-(n+N)T) \\ &= \sum_{n=-N}^{N} C_n W_{in}((k-n-N)T) \end{aligned} \qquad \text{Eq. 4.5}$$

According to the zero forcing algorithm described by Eq. 4.3, Eq. 4.5 should satisfy

$$\sum_{n=-N}^{N} C_n W_{in}((k-n-N)T) = \begin{cases} 1, & k=0 \\ 0, & k=-\mathrm{N},\ldots,-1,1,\ldots,\mathrm{N} \end{cases} \qquad \text{Eq. 4.6}$$

Eq. 4.6 can be changed to matrix form as

$$\mathbf{W_{out}} = \mathbf{W_{in}C} \qquad \text{Eq. 4.7}$$

where

$$\mathbf{W_{out}} = \begin{bmatrix} W_{out}(-NT) \\ \vdots \\ W_{out}(-T) \\ W_{out}(0) \\ W_{out}(T) \\ \vdots \\ W_{out}(NT) \end{bmatrix} = \begin{bmatrix} 0 \\ \vdots \\ 0 \\ 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix}, \qquad \mathbf{C} = \begin{bmatrix} C_{-N} \\ \vdots \\ C_{-1} \\ C_0 \\ C_1 \\ \vdots \\ C_N \end{bmatrix}$$

and

and

$$\mathbf{W_{in}} = \begin{bmatrix} W_{in}(-NT) & W_{in}(-(N\text{-}1)T) & \cdots & W_{in}(-3NT) \\ W_{in}((-N+1)T) & W_{in}(-NT) & \cdots & W_{in}((-3N+1)T) \\ \vdots & \vdots & & \\ W_{in}(NT) & W_{in}((N\text{-}1)T) & \cdots & W_{in}(-NT) \end{bmatrix}$$

To solve Eq. 4.7, the gain set $\{C_n\}$ can be calculated as

$$\mathbf{C} = \mathbf{W_{in}}^{-1}\mathbf{W_{out}} \qquad \text{Eq. 4.8}$$

In the practical circuit, it is not necessary to solve Equation 4.7. The equation can be solved by an adaptive approach. There are two modes in operating this adaptive equalizer. One is the training period. During this mode, a known sequence is transmitted and a synchronized version of this signal is generated in the receiver simultaneously. The equalizer compares the received version to the generated one and uses the errors between the two versions to adjust the gain set $\{C_n\}$. When the training period is completed, the adaptive equalizer is switched to the decision directed mode. During this mode, the data is received correctly with very high probability, so the system operates at a very low error rate.

Digital technology is used in the implementation of the equalizer. The input signal is sampled and quantized. The transversal filter is replaced by shift registers where the quantized input signal is stored. The tap gains $\{C_n\}$ are typically stored in the memory of a micro-controller. Multiplication, addition, and $\{C_n\}$ adjusting are performed by the micro processor [26].

Time domain equalization does not care about the channel frequency response. No matter how complex the channel is, the equalizer only needs to adjust the impulse response of the channel to be zero at the sampling point. If the number of the tapped delay elements is large enough, the intersymbol interference can be reduced to a sufficiently small amount.

Generally time domain equalization has advantages over frequency domain equalization. But the time domain equalizer is much more complex and costly than the frequency domain equalizer. Therefore, frequency domain equalization still has its value in some applications.

4.4 Eye Diagram

The eye diagram (or eye pattern) is a practical method to see the intersymbol interference. In the laboratory, the eye diagram can be recorded by the system shown in Fig. 4.9

A pseudo-noise (PN) signal generator generates a pseudo random data sequence. The data is encoded to a certain channel code by the channel encoder and transmitted over the channel. The circuit details of the PN signal generator and the encoder are shown in Appendix 2.

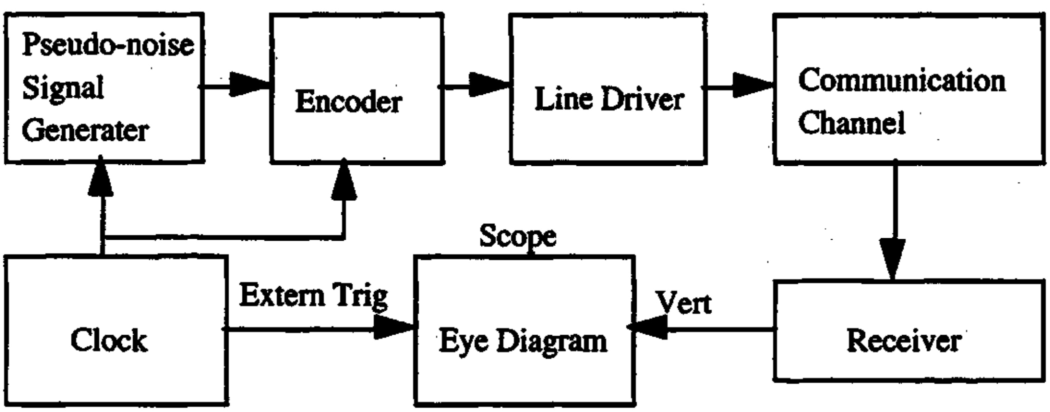


Fig. 4.9 System for laboratory eye diagram recording

The received signal is the input to the vertical deflection plates of an oscilloscope and the data clock signal is used to trigger the scope. Thus, the scope overlays equal length segments of the received waveform. The resulting display looks like human eyes, so the result is called an eye diagram.

The eye diagram of the biphase code which is used in LocalTalk is shown in Fig. 4.10. There are two smaller eyes in one big eye. After the data signals are transmitted through the communication channel, the original data eye diagram (see Fig. 4.10) is distorted by the intersymbol interference and noise (see Fig. 4.11). Under normal operating condition, the eyes are open. If there is a great deal of intersymbol interference or noise, the eyes will close. This indicates that a lot of bit errors should occur at the receiver.

The eye diagram provides an excellent way of analyzing the quality of the received signal. The interior region of the eye diagram is called the eye opening. The width of the eye opening (W in Fig. 4.11) defines the time interval over which the received wave can be sampled without an error from the intersymbol interference. The best time for sampling is at the point where the vertical opening of the eye is the largest. The sensitivity of timing error is given by the slope of the open eye and the noise margin of the system is given by the height of the eye opening (h in Fig. 4.11) [19]. Larger and clearer eyes have less errors and higher noise immunity.

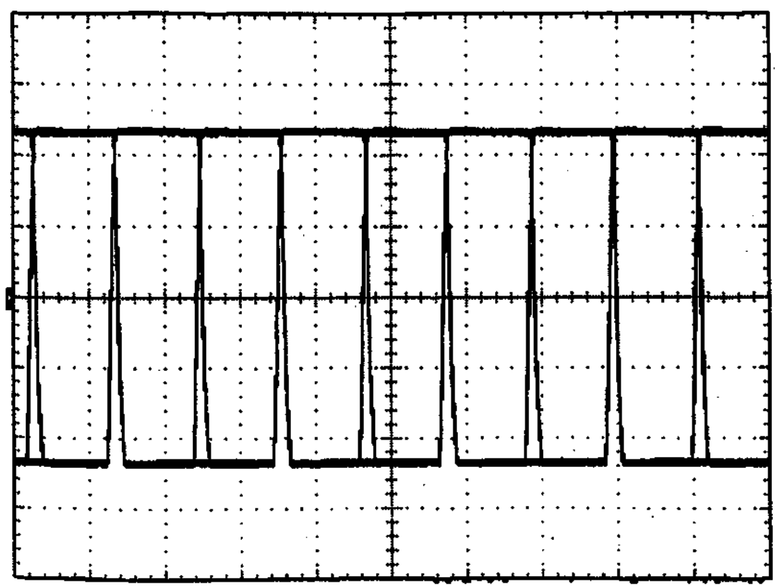


Fig. 4.10 Eye diagram of the undistorted data signal

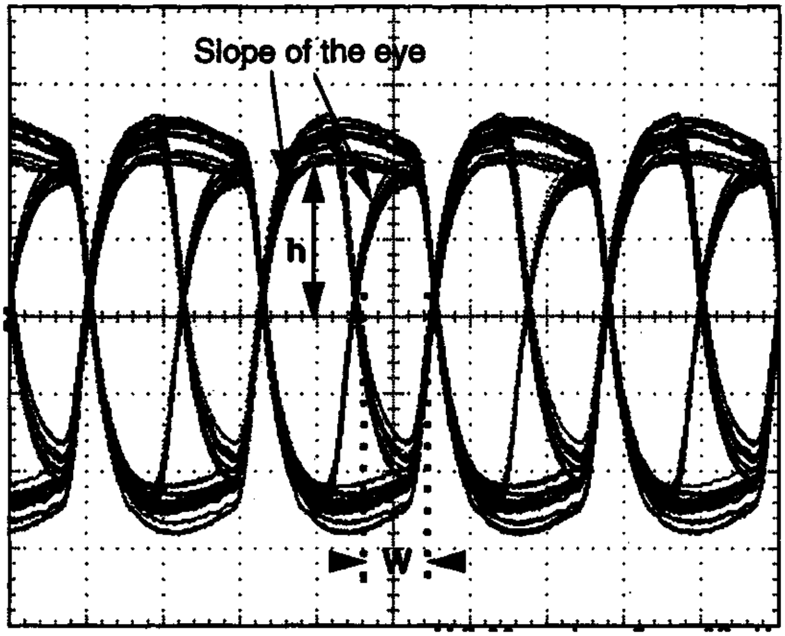


Fig. 4.11 Eye diagram of the distorted data signal

Chapter 5 : Voice and Data Simultaneous Communication

There are problems when transmitting unfiltered LocalTalk data and voice simultaneously over a subscriber line. The telephone user can hear noise caused by the data transmission. Also, the voice or address signaling in the telephone network causes errors in the data transmission. To realize voice and data communication simultaneously, the interference problem between data and voice must be solved.

Before further discussing the solutions to the interference problem, we need first to examine the characteristics of the signals transmitted on the subscriber line. After understanding the signals transmitted on the subscriber line, we have to investigate why this problem occurs, how serious it is and find a solution for this problem.

5.1 Signals Transmitted on the Subscriber Line

In the existing telephone network, the subscriber line can transmit voice, subscriber address signaling and in-band modulated digital signals (such as fax). In addition, the subscriber line will have to carry the LocalTalk data according to this specific project.

The main function of the subscriber line is to transmit voice signals. The full frequency range of the voice signal which can be heard is 20 Hz to 20 kHz. However, the power of the voice is mainly concentrated on the lower frequencies. Thus, in order to achieve a fairly good voice quality with less cost, the telephone network is designed to transmit voice frequencies that are lower than 4000 Hz. Traditionally, the average voice level in telephone communication is measured with a VU (volume unit) meter. Studies show that the output of the telephone set is about -13 to -15 VU with the average talker. Because the volume unit meter is a special meter with specified ballistic characteristics, the VU has no direct relation to dBm and it is hard to compare it to other measurements in dBm. In order to quantify the voice level in dBm, two new measurements of voice signals are used [27]. They are called equivalent peak level (EPL) and long-term conversational level. Both of them can be measured in dBm. Tests indicate that the average EPL at 0 transmission point is about -12 dBm and the average long-term conversational level is about -27 dBm.

In-band modulated digital signals are another type of signal that the subscriber line has to carry. The telephone network treats the in-band modulated digital signals as voice, but there are strict specifications for the transmission levels of this type of signal. If the signal levels are too high, they will cause problems to the network and decrease the telephone network service quality. The maximum allowed modulated signal level for the data terminal is -9 dBm.

Signaling is necessary to make terminals that are connected on the subscriber work properly with the telephone switch. Signaling can be classified as subscriber originated or central office originated signaling. There are two subscriber originated signaling types. The first one is to initialize a call by seizing the line. This is operated by taking the telephone handset off hook which places a 200 Ω or so impedance across the phone line. As soon as the central office senses the dc loop current caused by the off hook, it will assign some necessary resources to this subscriber and send dial tone to the telephone. The second subscriber originated signaling type is dialing. After dial tone is received by the subscriber, the destination phone number can be sent to the central office. There are two kinds of dialing methods. One is dial pulses. The pulses are produced by opening and closing the subscriber loop. The normal pulse rate is 10 pulses per second. Another dialing method is dual tone mulifrequency (DTMF). Each push button on the touch tone telephone is represented by two tones. The frequencies of the tones range from 697 Hz to 1633 Hz, with power varying from -14 dBm to +4 dBm.

The central office oriented signaling is used to control or inform the subscriber terminal. The most often used signaling is listed in Table 5.1

Signaling	Frequency	Level
Dial Tone	350 + 440 Hz	-13 dBm
Busy Tone	480 + 620 Hz	-24 dBm
Ringback Tone	480 Hz	-19 dBm
Ring	16 Hz to 66 Hz	37 dBV

Table 5.1 Central office originated signals

Dial tone is sent to those subscribers who are ready to dial the phone number and tell them that the central office is ready to receive dialing numbers. A busy tone indicates the called party is busy. When the called party is ringing, the calling party receives a ringback tone. Except the ring, the power of this type of signaling is low. The ring has the highest power but the lowest frequency.

5.2 The Interference Between Voice and Data Communication

The LocalTalk data, which has been discussed in Chapter 2, is a balanced biphase space data. It has 5 V amplitude with a special spectrum. The power of the data is mostly concentrated on the frequencies that are higher than 30 kHz.

In order to investigate interference between voice signals and data signals, a Macintosh (Mac) computer is connected to a subscriber line by a test LocalTalk interface module shown in Fig. 5.1.

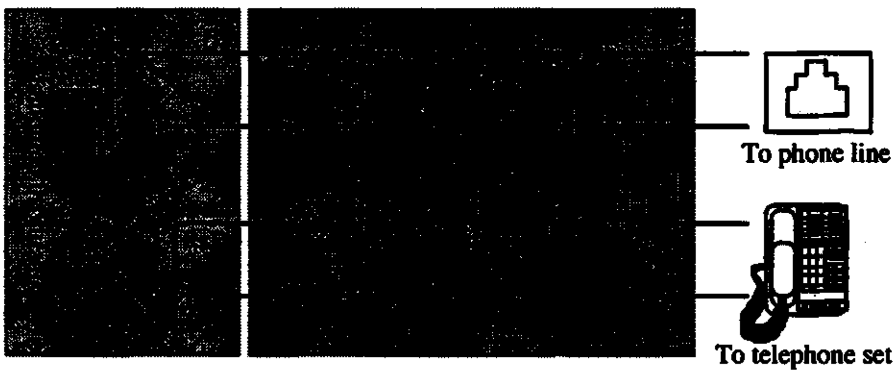


Fig. 5.1 A simple test interface module between a Mac and a subscriber line

The LocalTalk data, which comes from a Macintosh computer, is transmitted as balanced signal TxD+ and TxD- by the driver. The outgoing balanced LocalTalk data is connected to a subscriber line at J1 through a pair of large value capacitors. J2 is connected to a telephone set. The incoming signals RxD+ and RxD- are connected to the balanced receiver in the Macintosh computer.

The first experiment is to measure the data noise in the voice band. A unit called dBrnC is used. dBrnC is the noise power weighted with a C-message weighted filter and compared to a 1 pW reference noise which has -90 dBm level. The C-message frequency weighting curve is shown in Fig. 5.2.

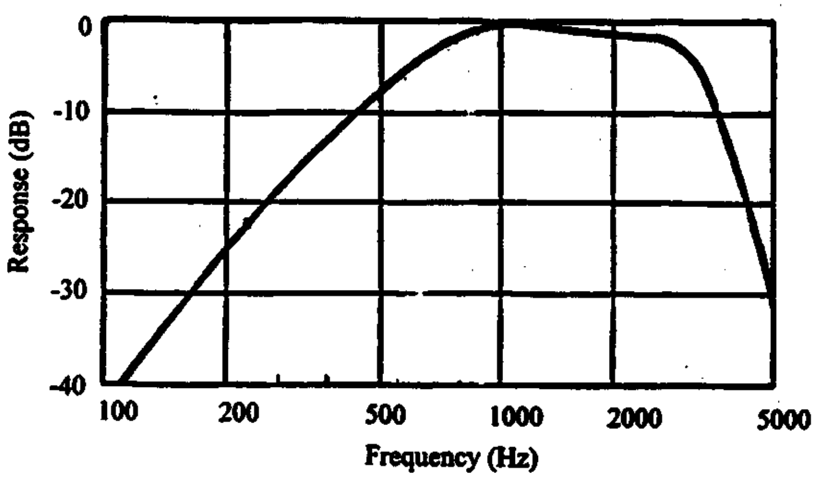


Fig. 5.2 C-message weighting curve [21]

Studies show that, at low sound amplitude, the hearing sensitivity of a human ear varies with frequency. Human ears are most sensitive to frequencies around 1000 Hz. Therefore, noise signals whose frequencies are in this range are most disturbing. The C-message weighting curve is based on the frequency response of a telephone set and the hearing sensitivity of human ears. This weighting curve can reflect the difference of the disturbance that noise signals cause to human ears. For example, according to the weighting curve, a 100 Hz noise signal with 0 dBm power is as disturbing as a 1000 Hz noise signal with level –40 dBm.

The data noise is measured by a SAGE 930A communication test set. This test set can not only measure the important parameters of the telephone network, but also can simulate a central office switch to supply dc power to the loop and support most subscriber loop signaling. In order to record the data noise that a telephone user can hear, the test set, instead of the telephone, was connected to J2 in Fig. 5.1. According to the specification of the communication test set, the noise measurements should be conducted with the far end of the subscriber line terminated by a 600 Ω or 900 Ω resistor in series with a 2.16 microfarad capacitor [28]. While the computer was sending LocalTalk data packets, the data signal was measured at about 81 dBrnC noise in the voice band. However, the noise caused by an idle telephone circuit is allowed to be 23 dBrnC or less. Therefore, the data noise is much greater than the idle telephone circuit and the data noise can be heard clearly.

The noise power in dBrnC can be expressed in dBm by the following relation

$$N_{dBm} \geq N_{dBrnC} - 90 \text{ dBm} \qquad \text{Eq. 5.1}$$

Acording to Eq. 5.1, the power of 81 dBrnC noise is larger than -9 dBm. Such high power of the noise can seriously disturb voice band fax communication. For example, if a subscriber line is used to transmitted 14.4 kbps fax data, a minimum S/N (signal to noise) ratio of 28 dB is required [28].

According to specifications, the maximum output power of a fax is -9 dBm and in this case

$$S/N = \frac{S_{dBm}}{N_{dBm}} = \frac{\leq -9\ dBm}{\geq -9\ dBm}$$

$$S/N \leq 0 \qquad \text{Eq. 5.2}$$

Obviously a fax machine can not work properly under this condition.

Why is there so much noise in the LocalTalk data? In Chapter 2, the power spectral density of a random infinite biphase space binary sequence are discussed. In theory, there is no dc component and very little power in the voice band. However the LocalTalk data packets are not infinite; some of them are quite short. The RTS and CTS control packets which are sent before each data packet are only 72 bits long. The probability of each bit pattern is no longer equal and random. There are four types of bit patterns in a LocalTalk data stream, which are shown in Fig. 5.3.

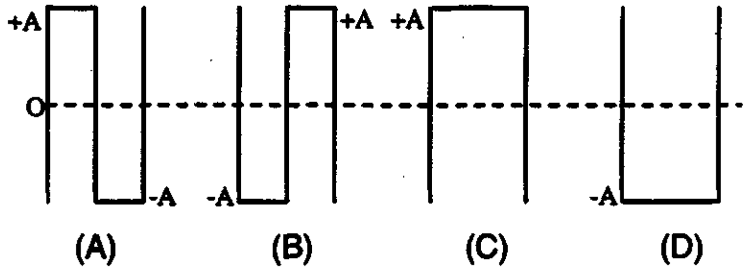


Fig. 5.3 Bit patterns of the biphase data

Each "0" pattern (A or B) is a bipolar symmetric pulse and does not produce any dc component. If the number of "0" patterns is not equal to an even number, they will not cause any extra noise in the voice band. But if "1" patterns (C and D) are not even in one of the packets, an extra positive or negative pulse will occur. It can result in dc and low frequency power in the packet. Even if the number of "1" patterns have an even number in one packet, The low frequency energy is still enhanced if they are separated by several bit times. Another reason that the LocalTalk packets generate a lot of voice band noise is due to the synchronization pulse. It is a single unipolar pulse (see Fig. 2.16), added to the front of each packet. If the sync pulse occurs every 600 μs (IDG+IFG), the frequency is about 1600 Hz, which results in large energy in the voice frequency range. Finally the output of the transmitter may not be equal on the positive half and the negative half. This unbalanced amplitude will result in voice noise as well.

Another experiment is to test the voice interference to data communication. The connection of the test system is the same as that of the last experiment. The Macintosh computer keeps transmitting data and the test set generates a 1000 Hz tone. The experiment result shows that only if the power of the tone is decreased below -25 dBm, will it stop disturbing the data communication. Unfortunately, the power of voice band signals which are normally transmitted on the subscriber line are much larger than -25 dBm and would therefore interfere with data transmission.

5.3 Proposed Solutions for Voice and Data Simultaneous Transmission

The objective of a solution to the interference between data and voice is to isolate voice and data signals and make them transmit on the same medium without sacrificing the service quality offered by each of them separately.

One proposed method for reducing the data noise in voice band is to reorganize the LocalTalk data. According to the causes of the data noise, the sync pulse could be substituted by a "0" pattern (A or B in Fig. 5.3). This type of sync pulse can play the same role as the '1' patterns sync pulse do, but have the advantage of no dc component. A bit stuffed method could be used to make "1" patterns in a LocalTalk data packet be even and not be separated too far away. This method can reduce the dc and low frequency components in a LocalTalk data packet. However, these methods have to be realized at the data link layer, therefor the original LocalTalk design must be changed. Obviously, this method is relatively complex, inflexible and costly.

Another approach, which is similar to frequency division multiplexing, is illustrated in Fig. 5.4.

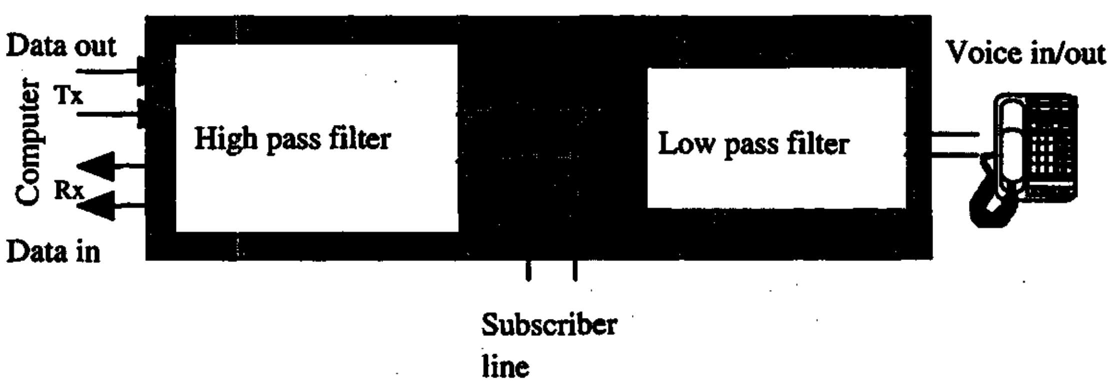


Fig. 5.4 New LocalTalk interface module for data and voice simultaneous communication

In this proposed method, a new LocalTalk interface module is designed. There is one high pass filter and one low pass filter in this module. Both the computer and the telephone set are connected to a subscriber line via this interface. The transmitted data (Tx) from the computer have to pass the high pass filter before being transmitted on the subscriber line and received signals from the subscriber line also have to go through the high pass filter before being received by the computer. The signals communicating between the telephone and the subscriber line are filtered by the low pass filter.

The high pass filter is used to restrict the LocalTalk data to the high frequency band and to reduce the noise of the LocalTalk data in the voice band. Although the real LocalTalk data packets generate noise in the voice band, their main power is still distributed on the frequencies that are higher than voice band. Therefore, if the high pass filter is designed properly, the voiceband power of the transmitted data can be remarkably reduced without losing the main information needed for the receiver decoding of the data. The high pass filter can also prevent voice band frequencies from interfering with the received data signals, since the high pass filter only allows high frequencies signals passing to the computer. The low pass filter also plays two roles here. One is to prevent high frequency data signals from reaching the telephone. Another role is to restrict the voice signals transmitted from the telephone to the low frequency band. Because of these filters, voice signals and data are transmitted in the same cable but in different frequency ranges

Using this new interface module, the original design of the computer and the telephone set do not need to be changed. Therefore, this method is very simple, flexible and cheap for application. These advantages are what this project needs.

5.4 The Circuit Design of the New LocalTalk Interface Module

The key issue of the new LocalTalk interface circuit is to design filters. The objective of designing these filters is to separate LocalTalk data and voice signals to different frequency ranges which can prevent interfering with each other. The new LocalTalk module with highpass-lowpass filters has been developed by Greg Erker and Garth Wells of TRLabs. Design criteria and performance measurements are presented in this section as background for the work described in Chapter 6.

Experiment has shown that LocalTalk data packets have up to 81 dBrnC noise on the voice band, if the data packets are transmitted to a subscriber line without any filters. In order to make the data noise quieter than the telephone noise, the high pass filter, working with the low pass filter, should reduce the data noise to less than 23 dBrnC. On the other hand, for the transmitted LocalTalk data, the pass band of the high pass filter should start at about 20 kHz which is the highest frequency that human ear can hear. We expect that the high pass filters can have as high attenuation as possible to the frequencies that are lower than 20 kHz. Frequencies that are higher than 30 kHz carry the main power of the LocalTalk data, therefore, we do not want the high pass filter to give too much attenuation to these high frequencies. When receiving data, the high pass filter should have enough attenuation to the voice band signals for preventing the interference to the received data signals. The pass band of the filter should still start from 20 kHz or so to minimize the distortion to the received data. The alerting or ringing signal has very high power but very low frequency, so the high pass filter should also be able to reduce the power of the ringing signal to a very low level.

Since 4000 Hz is the highest voice band frequency on the telephone network, the low pass filter should have pass band up to 4000 Hz and give as much attenuation as possible to the frequencies that are higher than 4000 Hz.

The circuit of the new LocalTalk interface module is shown in Fig. 5.5.

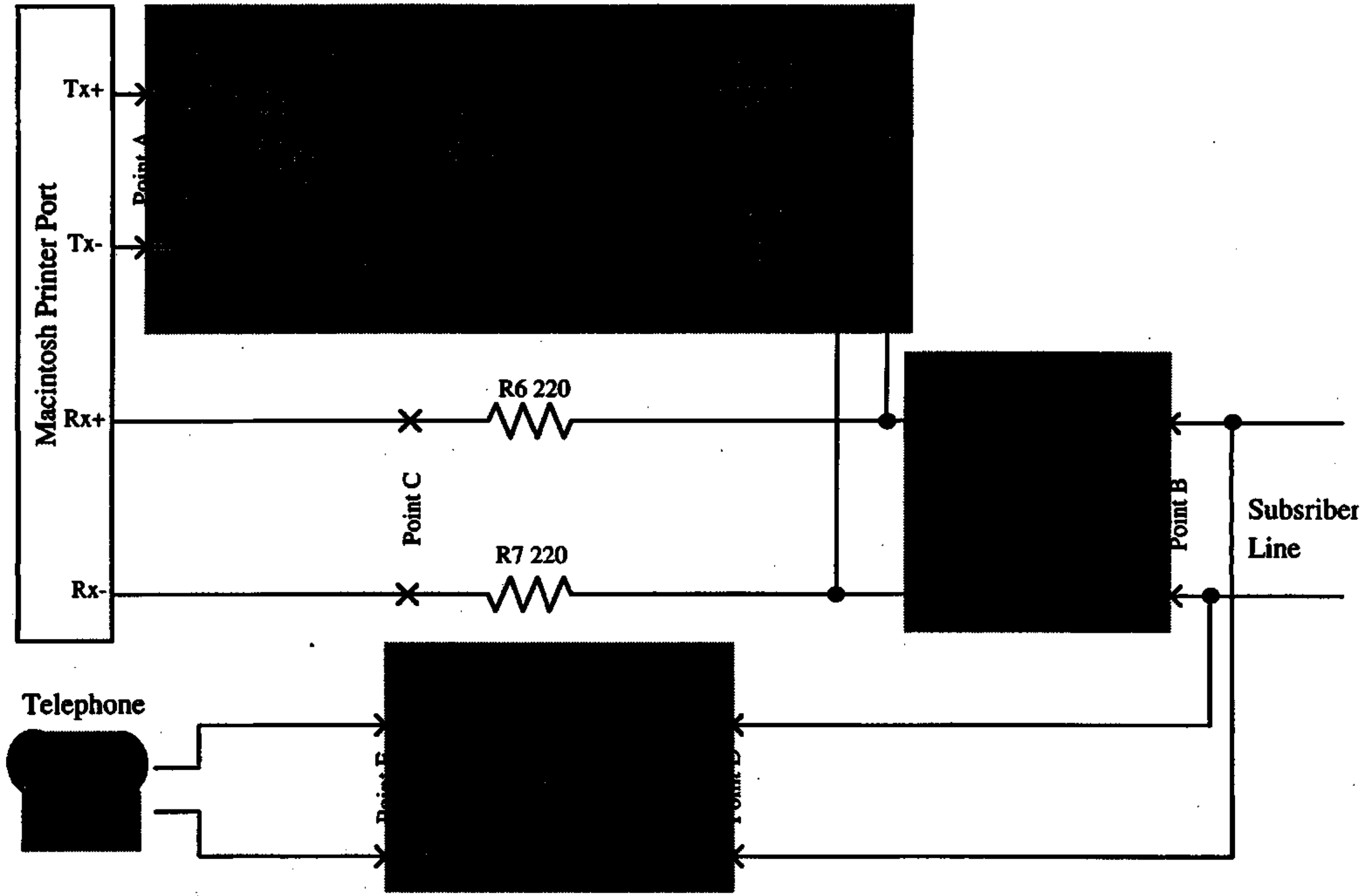


Fig. 5.5 New LocalTalk interface module

The high pass filter shown in Fig. 5.4 is separated to high pass filter A and high pass filter A1 (see Fig 5.5). LocalTalk data at Tx+ and Tx- from the Macintosh computer have to go through the high pass filters before being transmitted to the subscriber line. To reduce the amplitude of the transmitted LocalTalk data, a 4 to 1 ratio transformer is used. The incoming data is directly connected to the Rx- and Rx+ in the Macintosh printer port via R6 and R7.

The frequency response of the high pass filter A and A1 to the outgoing data (from point A to point B in Fig. 5.5) is shown in Fig. 5.6.

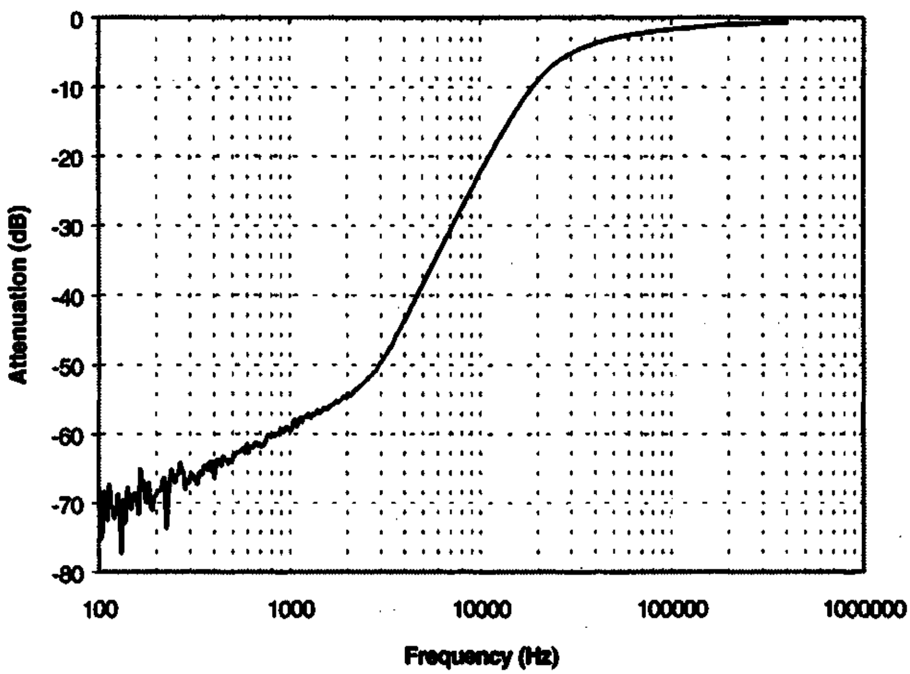


Fig. 5.6 The response of the high pass filter when transmitting data

The frequency response of the high pass filters give the LocalTalk data at least 40 dB attenuation to the frequencies that are lower than 4 kHz. Such high attenuation can reduce the data noise to a very low level in this band. The frequencies that are higher than 30 kHz have relatively flat frequency response and less attenuation. Such frequency response can minimize the distortion to the data by maintaining the high frequency power relatively unchanged. Fig. 5.7 shows the LocalTalk data before and after the high pass filters.

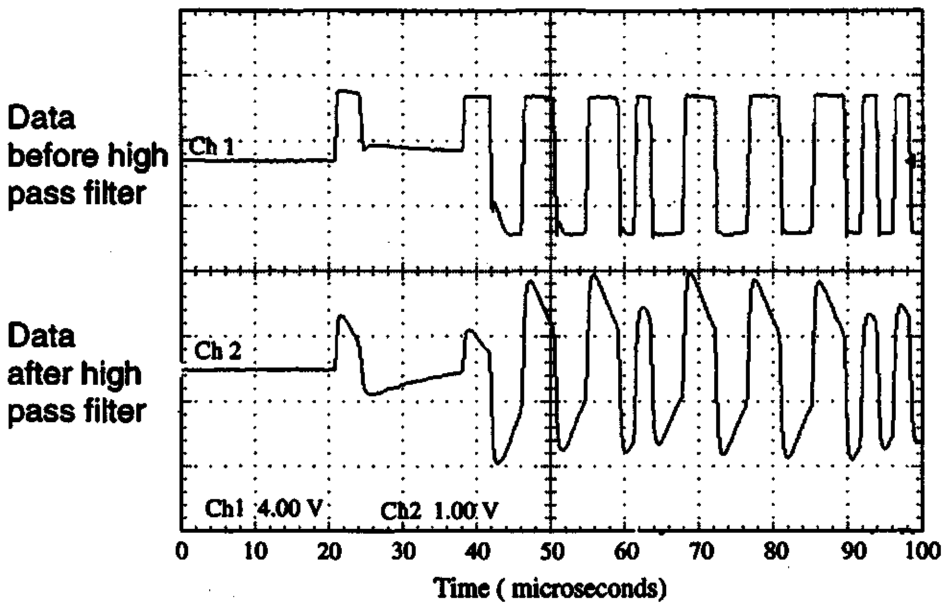


Fig. 5.7 LocalTalk data before and after the high pass filter

Fig. 5.7 shows that the shape of each pulse in the original data (before high pass filters) is changed to pulses that have a slight decrease curve (see the data after high pass filters), which has less low frequency components. However, the width and the transition point of each pulse, which are very important for receiver to decode the signal, are unchanged. Therefore, this distortion has no harm to data communication.

There still exists some data noise at frequencies above the voiceband. This high frequency data noise is due to the lower attenuation of the high pass filters to the frequencies ranging from 4000 Hz to 20 kHz. The low pass filter can help to eliminate this noise in the station set.

The low pass filter shown in Fig. 5.5 is composed by two 10 mH inductors (L2 and L3) with the load of an offhook telephone set. The frequency response of the low pass filter (between point E and D in Fig. 5.5) is shown in Fig. 5.8.

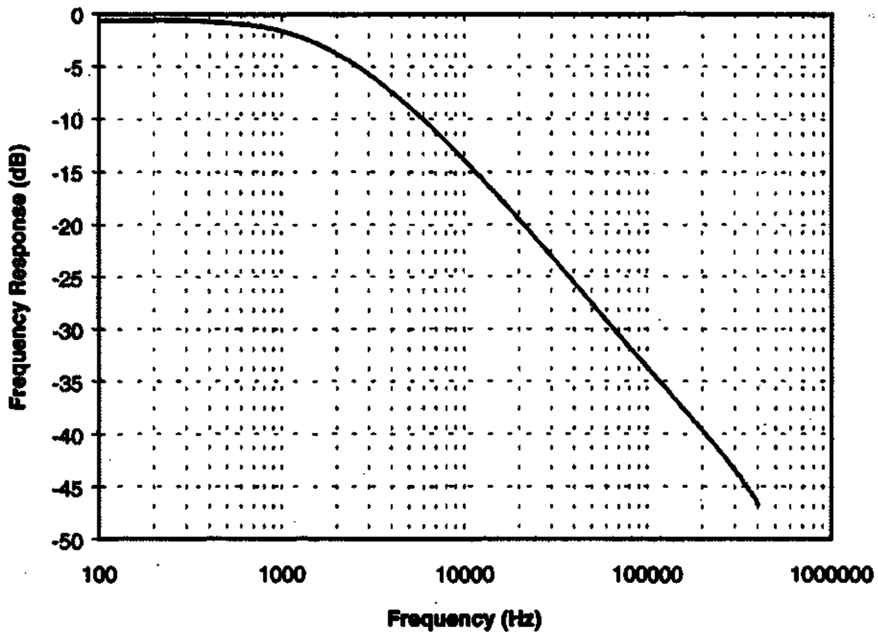


Fig. 5.8 The response of the low pass filter

The response curve has less than 3 dB attenuation to the frequencies that are lower than 2000 Hz, so this low pass filter shouldn't cause any troubles to the voice communication. The attenuation to the high frequencies is about 17 dB per decade. The frequency response of the low pass filter combined with the high pass filter, shown in Fig. 5.9, gives the transfer response from the data signal to the telephone receiver. Frequencies lower than 20 kHz are attenuated by at least 35 dB and frequencies less than 4 kHz are attenuated by at least 50 dB. This response means that the telephone user can not hear any data noise.

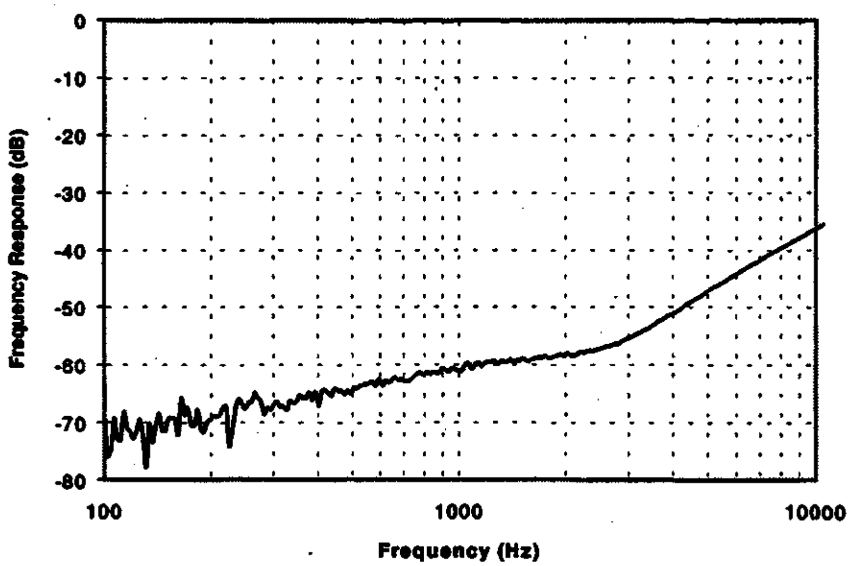


Fig. 5.9 The total attenuation of the data noise

Because the outgoing data and the incoming signals to the computer have different paths, the frequency response of the high pass filter A and A1 to the incoming data (point B to point C in Fig. 5.5) has different frequency response. The frequency response is shown in Fig. 5.10. This frequency response has less attenuation to the voice band signals, but the pass band still starts from 20 kHz or so.

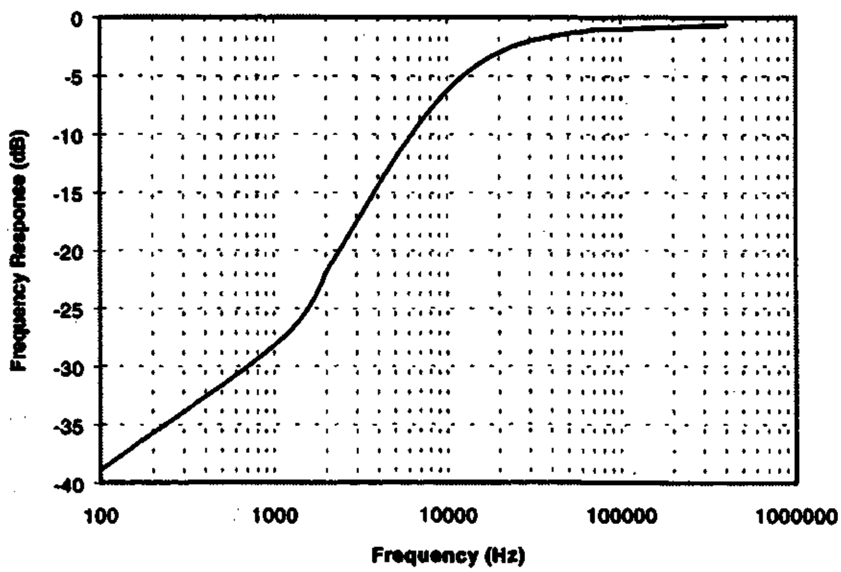


Fig. 5.10 The response of the high pass filter when receiving data

With helps of these filters, the new LocalTalk module can separate the voice and data into two different frequency channels. Some tests regarding the interference between the LocalTalk data and the voice with the new LocalTalk interface module will be presented in Chapter 6. Tests show that the LocalTalk data transmission has no disturbance on the voice communication and the data communication between two computers can perform well at distances up to 500 meters.

Chapter 6: Extended Range Transmission

Although the LocalTalk module developed in the previous chapter can realize voice and data communication simultaneously, its communication range is not long enough for application in the entire local telephone network.

6.1 The Local Telephone Network

The local telephone network is composed of central office switches and wire transmission links which connect subscriber equipment to the central office. An illustration of the local telephone network is given in Fig. 6.1. The feeder cable usually carries up to 3000 wire pairs. It starts at the central office and terminates at several drop off points called jumper wire interfaces (JWI). The area from JWI to the subscriber is called the distribution area. The length of the subscriber lines in the distribution area range from several hundred meters to 2 km. Very few cables are longer than 2 km [21]. If the new LocalTalk interface is to be used in the distribution area between the subscriber and the JWI, the communication distance must be up to 2 km.

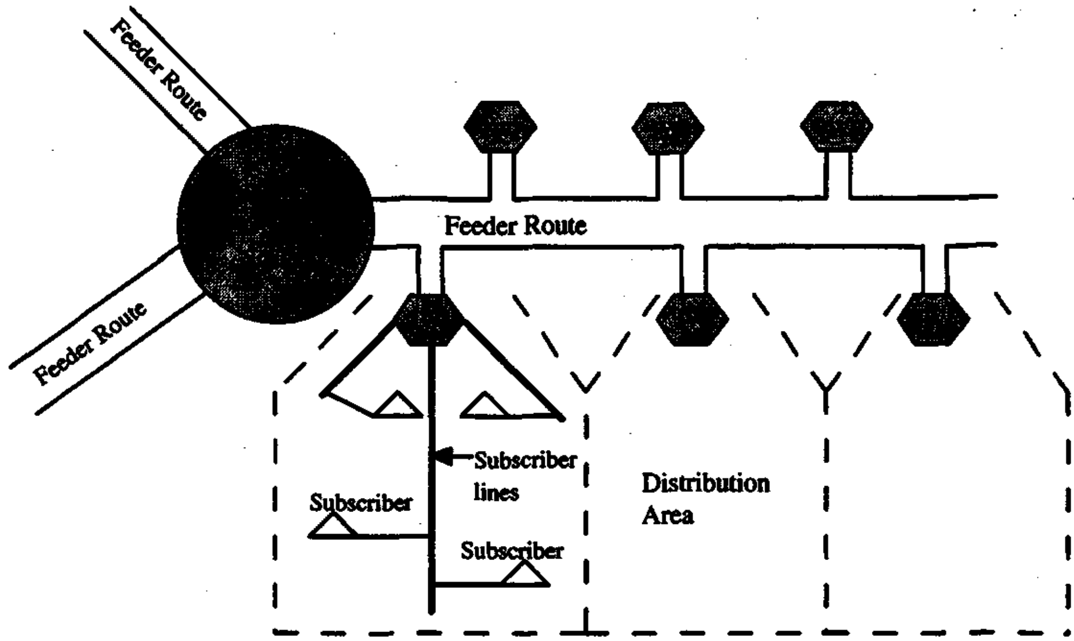


Fig. 6.1 Loop plant of the telephone network

In Chapter 4, the equalization and predistortion, which can extend baseband communication, were discussed. In this chapter, a practical equalizer, which is used in extending the communication distance of the new LocalTalk interface, will be presented. This along with the predistortion method will be discussed in this chapter. Finally the effect of bridged taps will be investigated.

6.2 Design of the Equalizer

In order to use the new interface module in the distribution area, the communication distance must be at least 2 km and the equalizer must be able to work at different distances and different temperatures. In this project, a low cost and high performance equalizer circuit is desired and the equalizer should not alter voice frequency transmission while supporting the simultaneous data and voice communication.

6.2.1 Equalizer Type

Time domain and frequency domain equalizers are mentioned in Chapter 4. The time domain equalizer has several advantages over the frequency domain equalizer. It uses DSP technology and an effective equalization algorithm to adjust the signal shape directly in the time domain. It has better performance in reducing intersymbol interference than the frequency domain equalizer and achieves better communication quality at higher speed and longer distance. However, this excellent performance is costly. The time domain equalizer is much more complex and expensive than the frequency domain one. In this project, the cost and the reliability are very important factors, thus the equalization method must be practical and efficient without sacrificing communication quality. The frequency domain equalizer is the first choice provided it has satisfactory performance.

6.2.2 Equalizer Circuit

Because the twisted pair channel, which is discussed in Chapter 3, has higher attenuation at higher frequencies and lower attenuation at lower frequencies, the equalizer should have higher gain at higher frequencies and lower gain at lower frequencies, which is the inverse characteristic of the twisted pair channel. A simple circuit is shown in Fig. 6.2

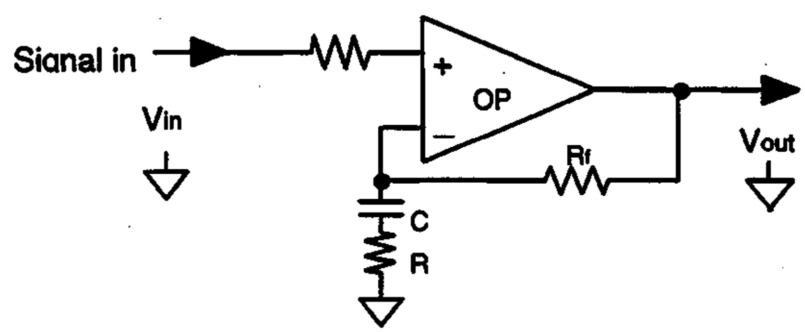


Fig. 6.2 Frequency domain equalizer

The operational amplifier gives necessary gain to the incoming signal. R, C and R_f can adjust the gain at different frequencies. The gain of this circuit is given by:

$$G = \frac{V_{out}}{V_{in}}$$
$$= 1 + \frac{R_f}{R + \frac{1}{j2\pi fC}} \qquad \text{Eq. 6.1}$$

where: V_{out} is the output voltage.
V_{in} is the input voltage.
f is the frequency.

By selecting component values, R = 110 Ω, R_f = 1100 Ω and C = 9500 pF, the theoretical equalizer gain is very similar to the inverse attenuation of a 2900 m AWG26 twisted pair. Fig. 6.3 shows the nominal gain of the equalizer and the inverse attenuation of the 2900 m AWG26 twisted pair.

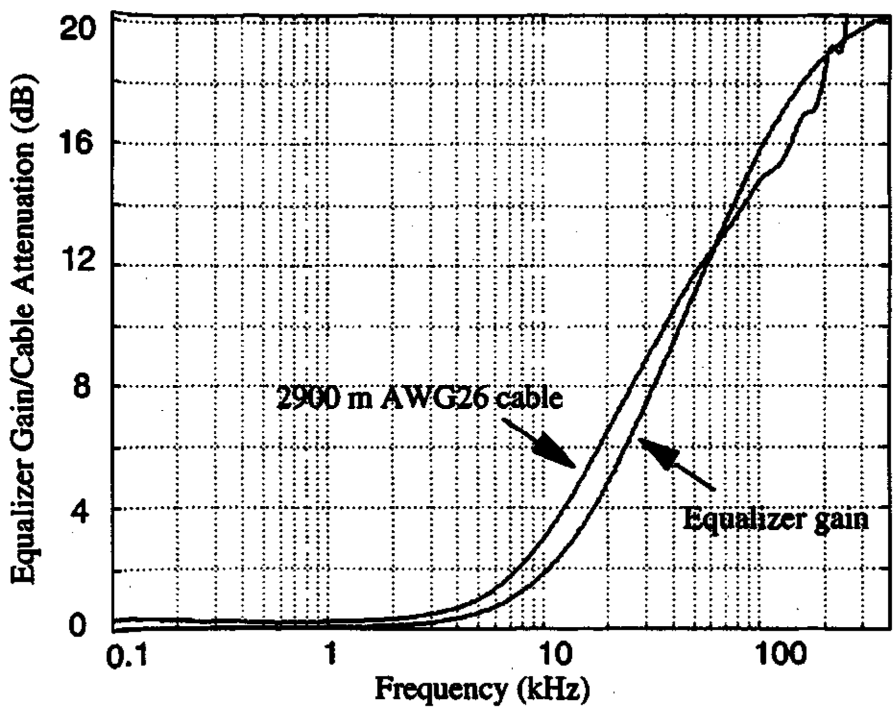


Fig. 6.3 Equalizer gain verse frequency

Although the curve of the equalizer gain is not exactly the same as the attenuation curve of the cable, the differences between the two curves are small at frequencies that range from 30 kHz to 300 kHz, in which the LocalTalk data has the highest signal energy. To achieve the exact inverse curve of the cable is difficult in the frequency domain and is not essential for accurate data transmission.

LocalTalk uses a balanced signal, so it is convenient to use one equalizer for each input line. The straight forward connection is shown as Fig. 6.4:

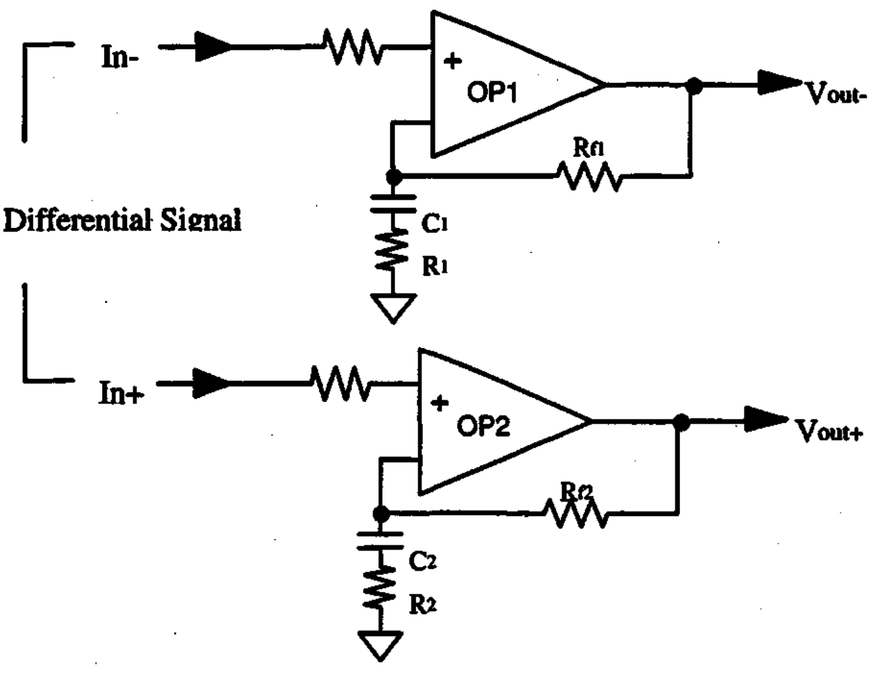


Fig. 6.4 Obvious equalizer for LocalTalk data

The fatal weakness of this circuit is that the balance of the equalizer can not be guaranteed. The balanced signal requires that the equalizer of each line have the same characteristic. Although the two operational amplifiers are quite similar and the values of resistors can be very accurate, the capacitance of capacitors can have as high as 20% errors over or below their labeled value. If the capacitance of C_1 has plus 20% and C_2 has minus 20% error, the unbalance frequency response of the circuit in Fig. 6.4 can go up to 1.5 dB from 30 kHz to 300 kHz. In this frequency range, LocalTalk data has the highest signal energy, so the unbalance of the equalizer certainly has negative effect to the equalized signal. With the increase of the transmission distance and the decrease of the signal to noise ratio, the negative effect of the unbalance of the equalizer is more notable.

An improved connection is shown in Fig. 6.5 [35].

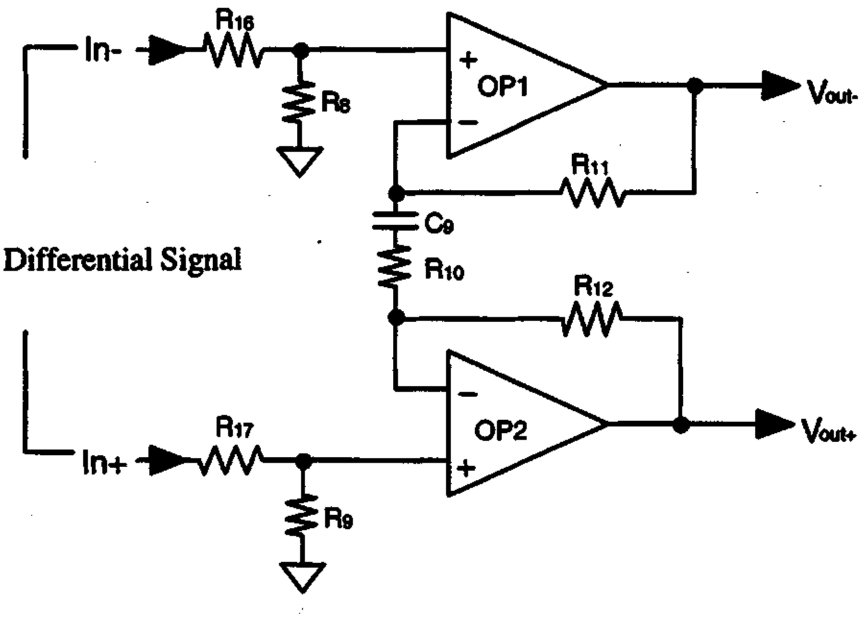


Fig. 6.5 The improved equalizer

R_{16} and R_{17} are used to protect the positive input of the amplifiers. R_8 and R_9 can give the input signals which are coupled via a transformer reference voltages to ground. The operational amplifier OP 1 and 2 share the same R_{10} and C_9. Any error of C_9 will cause the same frequency response change to both equalizers. Thus, the balance between the two channels still remains high. To maintain the balance of the circuit, R_{16}, R_8 and R_{11} must be equal to R_{17}, R_9 and R_{12} separately. In order to avoid decreasing the voltage of input signal, $R_8 \gg R_{16}$ and $R_9 \gg R_{17}$. The gain of this circuit can be calculated by

$$G(f) = 1 + \frac{2R_f}{\frac{1}{j2\pi fC_9} + R_{10}} \qquad \text{Eq. 6.2}$$

where $R_f = R_{11} = R_{12}$, f is the frequency.

If $R_f = 550\ \Omega$, $R_{10} = 110\ \Omega$ and $C_9 = 9500$ pF, the circuit has the same frequency response as Fig. 6.3.

The comprehensive circuit of the new LocalTalk interface with equalizer is drawn in Fig. 6.7 (see next page). The core component of the equalizer is the operational amplifier. It should have wide bandwidth and can handle the 230 kbits/s data rate. The TL082 integrated amplifier is selected for this project. It has unity-gain bandwidth of 3 MHz. Although the bandwidth will decrease while the gain increases, the bandwidth still has 300 kHz at gain of 20 dB [29]. The amplifier's slew rate is about 13 V/µs, which is adequate for the LocalTalk data [30]. The frequency response of the real equalizer is shown in Fig. 6.6

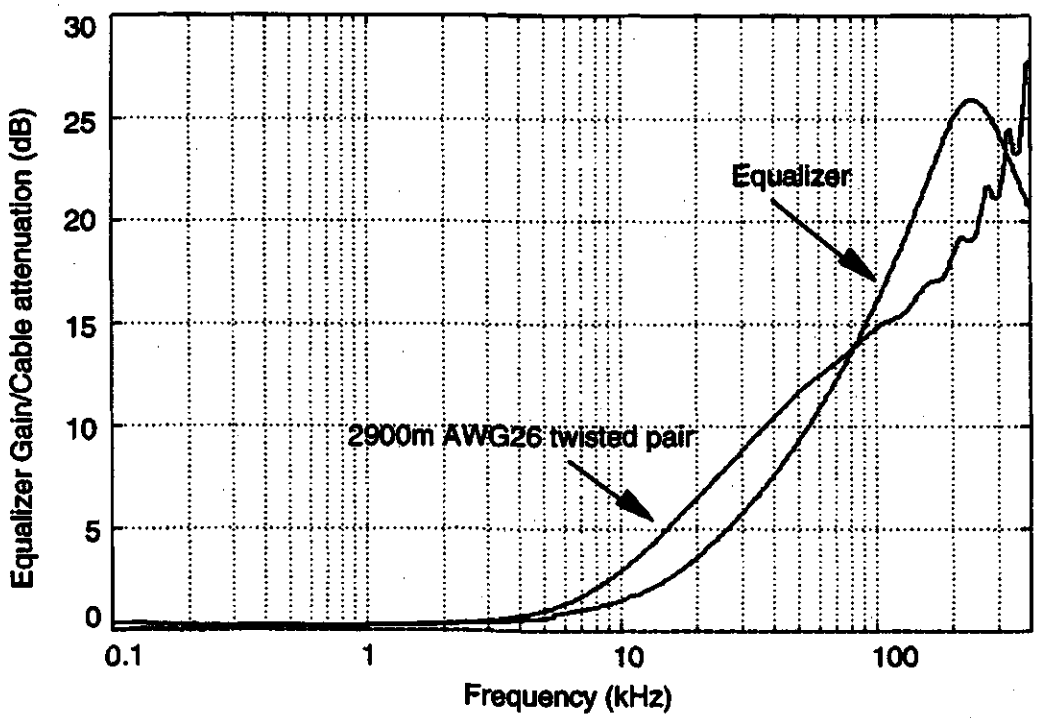


Fig. 6.6 Equalizer gain and cable attenuation verse frequency

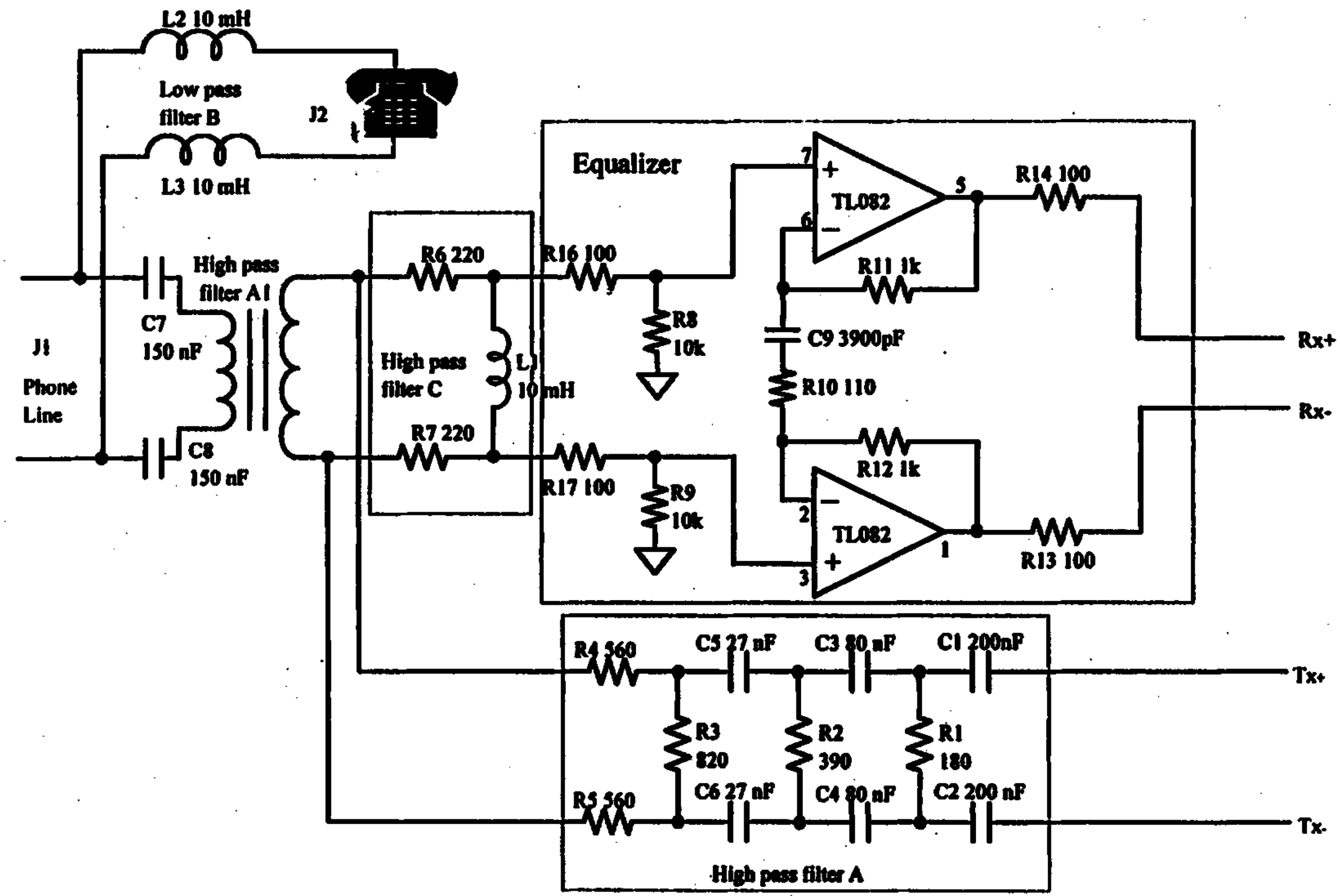


Fig. 6.7 New LoacalTalk interface with equalizer

The response shows that the equalizer has very little gain in the frequencies from dc to 10 kHz which is desirable in this project. The curve is still similar to the attenuation of the 2900 m AWG26 twisted pair. The gain of the equalizer at frequencies above 100 kHz is higher than the cable attenuation. It gives the '0' waveforms, which has more high frequency components and attenuates faster than the '1' waveforms, a strong compensation. Clearly the equalizer can not recover the data signal without distortion, but it can recover the important timing information as well as the amplitude that is required by the receiver. The result of the equalizer lets the eyes of the received signal open widely with less timing jitter (see Fig. 6.13 in Page 92 and Page 93).

The function of the high pass filter A, A1 and the low pass filter B are described in Chapter 5. The high pass filter C gives the equalizer more attenuation at the lower frequencies (see Fig. 6.8). It helps to reduce the interference from the voice.

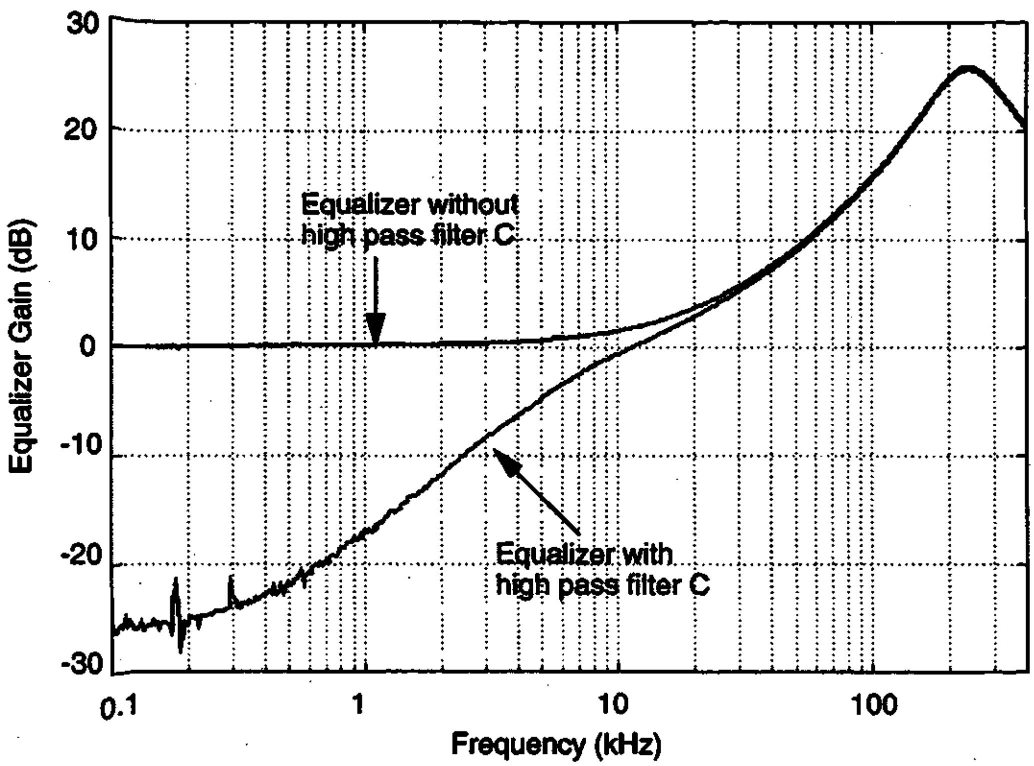


Fig. 6.8 Equalizer with high pass filter C

6.2.3 The Test Results of the New LocalTalk Interface with Equalizer

The equalizer is tested with respect to the data communication distance, data noise at voice band, telephone network signaling interference to data, and the circuit balance of the equalizer.

Circuit Balance of the Equalizer:

The block diagram for testing the circuit balance is shown in Fig. 6.9.

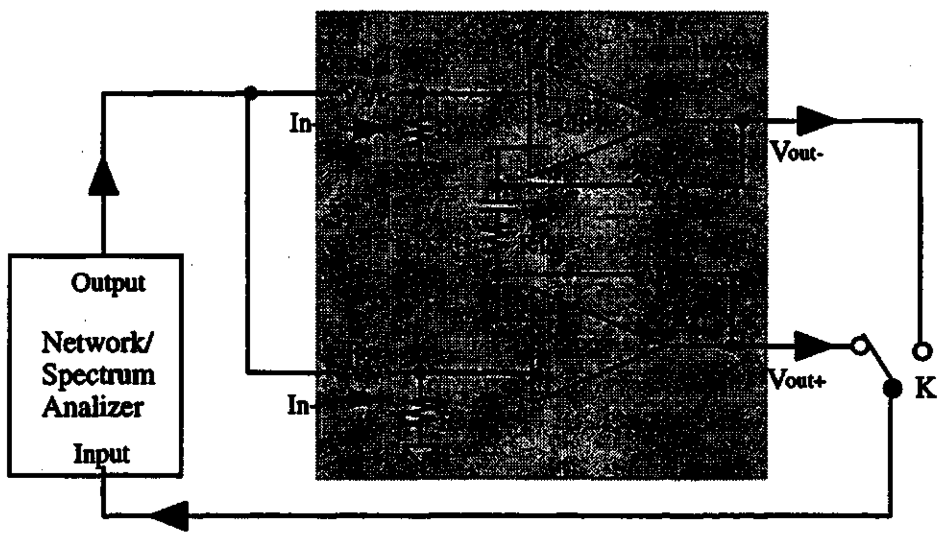


Fig. 6.9 Circuit balance test diagram

The sweep output of the network/spectrum analyzer is connected to the positive input of each channel's equalizer. Since the spectrum analyzer only has one input port, switch K is used to control which channel's output would be connected to the analyzer. The spectrum analyzer has to sweep the circuits twice to get the frequency responses for both channels. The unbalance of the two equalizer channels (shown in Fig. 6.10) is obtained by comparing the sweep result of each channel. The result shows that the unbalance of the two channels is well below 0.1 dB except three unbalance peaks. The three peaks are due to spurious noise which occurred during measurement. The circuit balance of the equalizer is acceptable.

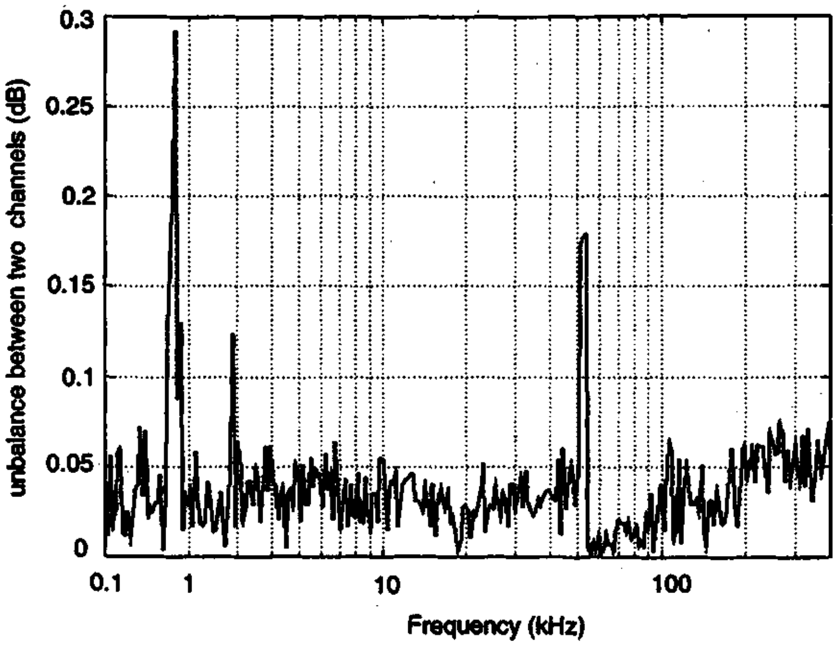


Fig. 6.10 The unbalance of the two equalizer channels

Data Noise at Voice Band

The data noise is still measured by SAGE 930A communication test set. The test system is shown in Fig. 6.11. The communication test set is connected to J2 of the interface module A and performs as a central office switch as well as a test instrument. dc power and the subscriber loop signaling are provided by this test set. This connection can make the AWG 26 twisted pair function as a subscriber loop and simulate that the new LocalTalk interface module is working on the real telephone network. The length of the twisted pair is about 715 m. According to the specification of the communication test set, the noise measurements are conducted with the far end of the subscriber line terminated by a 600 Ω or 900 Ω resistor in series with a 2.16 microfarad capacitor. This termination is called quiet termination and is connected to J2 of the new interface module B. Computer A can communicate with computer B via the new LocalTalk interface modules.

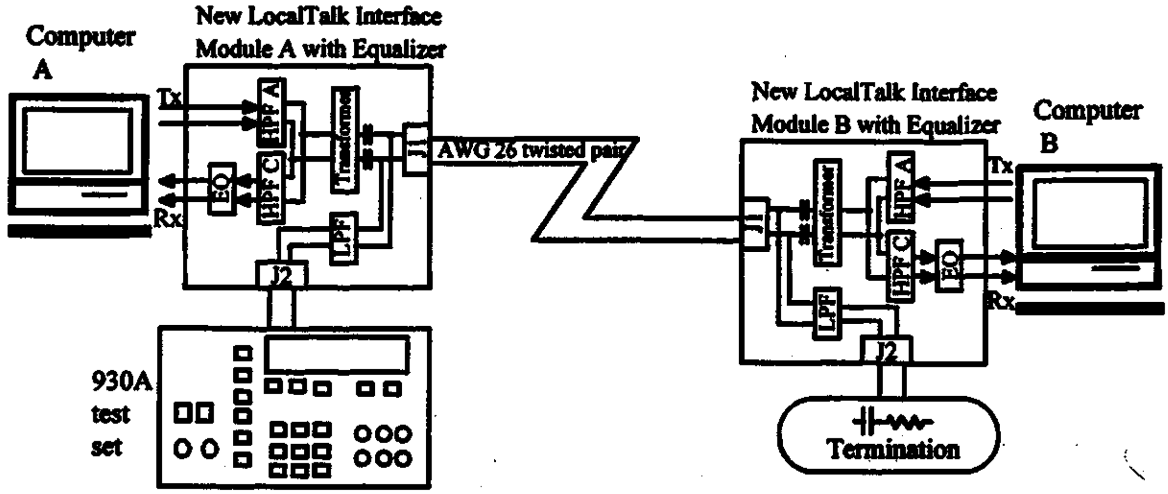


Fig. 6.11 Data noise measurement connection

When no data is transmitted on the twisted pair, the noise in the transmission loop is 0 dBrnC. If computer A and B continue sending data packets to each other, 4 to 6 dBrnC noise is measured. If we stop data transmission and substitute the quiet termination with a telephone set in the off-hook state, the noise is measured as 30 dBrnC. Obviously the noise caused by the telephone set on the subscriber loop is much higher than the data noise (30 dBrnC vs. 6 dBrnC), so the data noise can not be heard when using the telephone.

The signal to noise ratio (S/N) is measured by the same system shown in Fig. 6.11 except the quiet termination is substituted by a 1000 Hz tone generator. The test set treats the 1000 Hz tone as signal and other frequency components as noise. Experiments show that when the power of the tone is larger than -25 dBm, the S/N ratio has no difference with or without the data transmission. Therefore, the new interface module can not have any negative effects on the fax or other voice band data communication whose output level is usually higher than -25 dBm.

Data Communication without Telephone Signaling Interference

Two methods are used in measuring the quality of the data communication. One is to use the echo protocol in the transport layer. Computer A keeps sending echo data packets to computer B, while computer B receives a valid echo packet, it sends the packet back. The computer A can compare the number of echo packets being sent to the number of valid packets coming back, therefore, the computer A gets the percentage of dropped packets. The length of each data packet is 580 bytes and there are about 8000 packets being transmitted in each test case. If the dropped packets are high, this means that the transmission line or protocol has more errors.

Another method is to download a large random data file from computer B to computer A. A 473 kbyte JPEG compressed file was used for this purpose. The time needed to download this file is about 40 seconds by using AppleTalk protocols. A one to two second variation is treated as normal.

The system shown in Fig. 6.11 is still used in this test, but the cable length is varied from 0 m to 3360 m and no termination is necessary at J2 in module B. In the laboratory, an 11-pair AWG 26 cable roll is cut into two basic segments with lengths of 715 m and 1120 m. Several pairs in each segment can be connected to produce a longer pair. For example, two 1120 m pairs can be connected to a 2240 m cable.

The results of the dropped packets and download time at different lengths of cable are shown in Table 6.1.

Length of the cable	Dropped packets	Average down load time
0 m*	0.4%	38 seconds
715 m	< 0.05%	39 seconds
1120 m	< 0.05%	39 seconds
1430 m (2 x 715 m)	< 0.05%	39 seconds
1835 m (1120m +715 m)	< 0.05%	39 seconds
2145 m (3 x 715 m)	< 0.05%	40 seconds
2550 m (2 x 715 m +1120 m)	< 0.05%	40 seconds
2955 m (2 x 1120 m + 715 m)	< 0.05%	40 seconds
3360 m (3 x 1120 m)	< 0.05%	40 seconds

* 0 m means that the cable length is shorter than 2 m.

Table 6.1 The data communication quality at different cable lengths

The data shows that the new LocalTalk interface with the equalizer can work from very short distance to 3360 m.

Eye diagrams and waveforms at difference cable lengths are obtained by the system shown in Fig. 6.12. A similar system which is used to obtain eye diagrams has been described in Section 4.4, Chapter 4. The difference here is that the encoded signal need to be transmitted through the new LocalTalk module A and received via the interface module B. The eye diagrams recorded after the equalizer are obtained by connecting the output of the equalizer to the vertical inputs of the scope and the eye diagrams before the equalizer are recorded by connecting scope's vertical inputs to points a and b in the interface module B. Trace M1 in the scope is used to record the waveform of the received signal after the equalizer and the trace M2 is used to record the waveform before the equalizer.

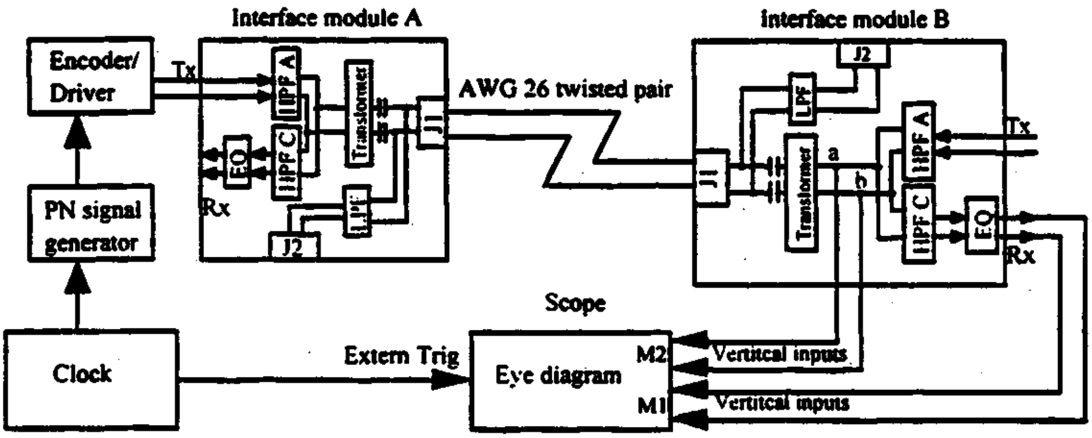


Fig. 6.12 The system to obtain eye diagram.

Fig. 6.13a and Fig. 6.13b show that the eye diagrams and waveforms which are recorded before and after the equalizer at different cable lengths. According to the waveforms in Fig. 6.13a and Fig. 6.13b, we can see that, at the same distance, the waveform recorded after the equalizer (trace M1) has better shape over the waveform recorded before the equalizer (trace M2) and at 0 meter distance, M1 becomes a square-wave, which means that the equalizer amplifiers are saturated by the large received voltage. According to the eye diagrams in Fig. 6.13a and Fig. 6.13b, we can easily determine that the eye diagrams after the equalizer open wider and have clearer border. Obviously the equalizer improves eye diagrams at every length.

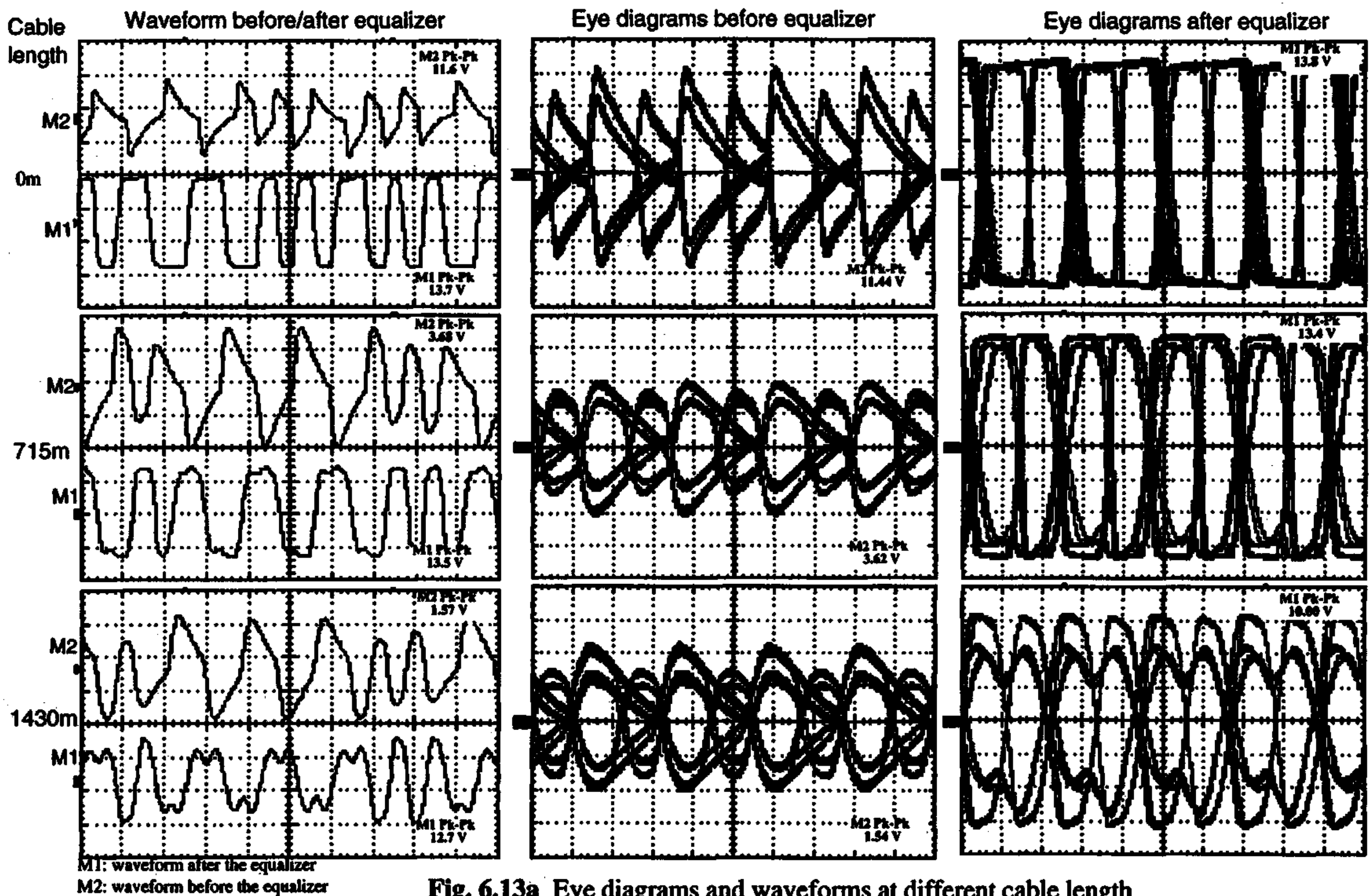


Fig. 6.13a Eye diagrams and waveforms at different cable length

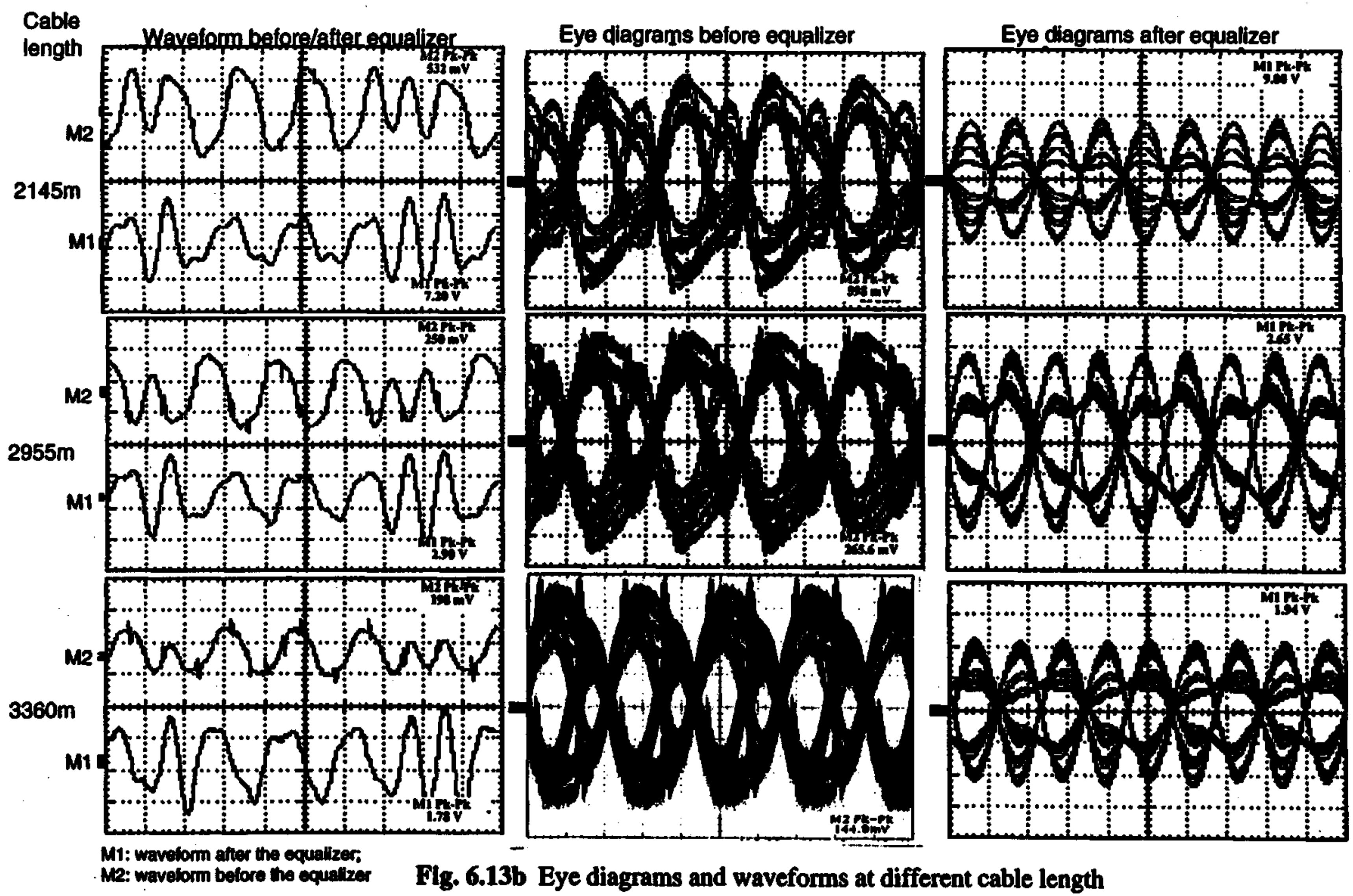


Fig. 6.13b Eye diagrams and waveforms at different cable length

Besides the equalizer, the high pass filter A and A1 (see Fig. 6.7) also helps to improve the communication distance. Generally, the frequency domain equalizer has difficulty in accurately equalizing the phase distortion, but the high pass filters help the equalizer recover some of this disadvantage. After the data goes through the high pass filters, the bandwidth of the data gets narrower and the spectrum of the data is restricted in the high frequency band. In Chapter 3, Fig. 3.5 and Fig. 3.7 show that the delay of higher frequencies has less variation, so the cables give less phase distortion to the data that goes through the high pass filter A and A1.

In order to illustrate the effects of the high pass filter A and A1, one of the experimental test systems is set up as in Fig. 6.14. This system can be used to record eye diagrams without high pass filter A and A1. Eye diagrams with high pass filer A and A1 are those unequalized eye diagrams which are shown in Fig. 6.13a and Fig. 6.13b.

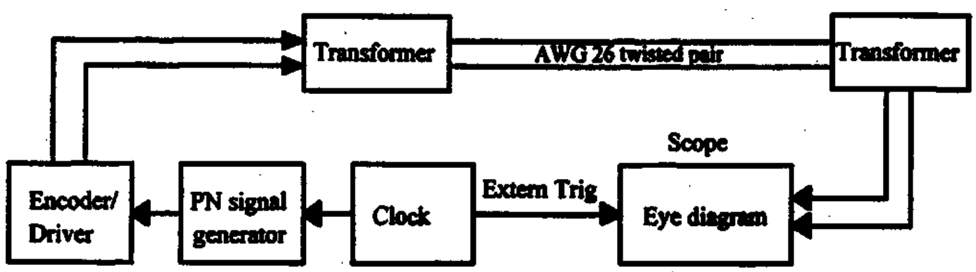


Fig. 6.14 Eye diagram recording system

Fig. 6.15 shows the effects of the high pass filter A and A1 to eye diagrams at different cable lengths. Eye A to Eye A_3 are eye diagrams without the high pass filters. Eye B to Eye B_3 are the eye diagrams with the high pass filters. At distances that are shorter than 1.5 km, Eye A and Eye A_1 have better shapes than those of Eye B and Eye B_1. However, with the increase of the cable length, the first small eyes of Eye A_1 get much smaller at 2200 m (see Eye A_2) and eventually disappear in Eye A_3 at 3300 m. On the contrary, at 2200 m and 3300 m, the two small eyes in Eye B_2 and B_3 are still open. Obviously the eye diagrams with the high pass filters are better at long distance.

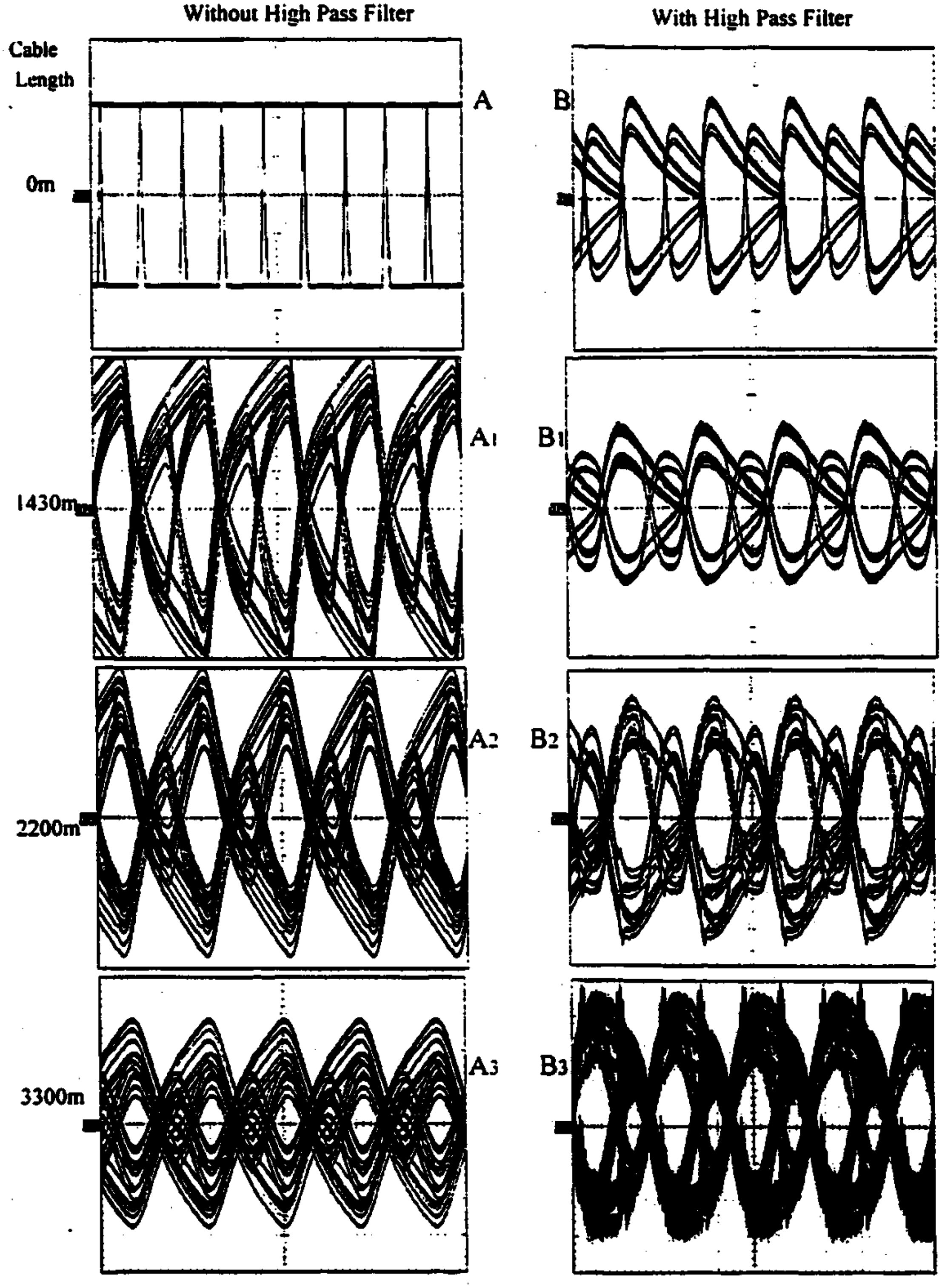


Fig. 6.15 Eye diagrams with/without high pass filter A and A1

Data Communication with Telephone Signaling Interference

The test system was again set up as Fig. 6.11 with substituting a telephone set in place of the termination. The SAGE 930A communication test set still takes the place of the central office switch. The dc power, test tones and some subscriber signaling are provided by the test set. The test results of dropped data packet rate at different distances are listed in Table 6.2.

Signaling/Cable length	715m	1430m	2240m	2955m	3360m
Ring supplied by SAGE 930A	0.0%	0.0%	0.0%	0.0%	0.0%
DTMF dialing by SAGE 930A and telephone set alternatively	0.0%	0.0%	0.0%	0.0%	0.0%
Pulse dialing by SAGE 930A and telephone set alternatively	0.3%~ 0.4%	0.2%~ 0.3%	0.3%~ 0.4%	0.2%~ 0.3%	0.2%~ 0.3%
1 kHz tone supplied by SAGE 930A; Output power = 12 dBm	0.0%	0.0%	0.0%	0.0%	0.0%

Table 6.2 Packet drop rate at different telephone signaling

According to the test results, ring and DTMF dialing have no effect to the data communication at each cable length (The drop rate remains 0.0%). When the 1000 Hz tone is added to the line by the test set, the output power of the tone can be as high as 12 dBm without any disturbing to the data transmission. Since the high pass filter C and the transformer in Fig. 6.7 have more attenuation at lower frequencies, it is very safe to say that most signaling tones, such as busy tone, dial tone and ringback tone, whose frequencies are lower than 1000 Hz and power is less than 12 dBm. have no negative effect to the data transmission. When pulse dialing, the phone set will switch itself back and forth from on-hook to off-hook. This causes impulse noise which has very large power and wide frequency range. The impulse noise causes some errors in the data transmission.

The effects of several higher frequency tones are also examined. There are no packets dropped at different cable lengths when a 1 dBm 2 kHz or a -7 dBm 3 kHz tone are added to the transmission line. This test result shows that the immunity of the data transmission to the voice band signals decreases with the frequency increasing. However, most of the voice power is concentrated around 1000 Hz. the voice signal's equivalent peak level is about -12 dBm and its long-term conversational level is only about -27 dBm. Therefore, the normal conversation does not interfere with the data communication.

6.3 Predistorter and Equalizer Design

Another possible way that data communication distance can be extended further is to use both predistorter at the transmitter side and equalizer at the receiver side. The diagram of the whole system is shown in Fig. 6.16.

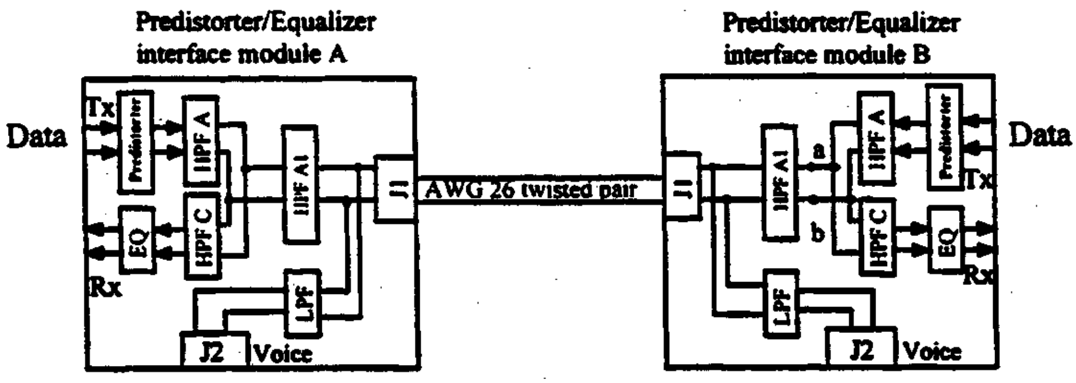


Fig. 6.16 System with predistorter and equalizer

Tx data comes from the computer. When the data goes through the predistorter, the data is distorted to a special shape which is suitable for longer distance transmission. At the receiver side, an equalizer is still used to compensate for the distortion caused by the transmission system. The high pass filter A, A1, C and the low pass filter perform the same roles as those described in Section 6.2. The following sections will describe the predistorter and the equalizer in detail.

6.3.1 Predistorter Design

The predistorter is based on the theory that a data signal whose spectrum is narrow and restricted at high frequency band has less phase distortion after long distance transmission, which is proved in Section 6.2.3 by using high pass filters. The method used in the predistorter is more effective.

The LocalTalk data distorted by the predistorter is illustrated in Fig. 6.17 C [31]. This distorted LocalTalk data has reduced amplitude of each later portion of each bit. The special shape of the distorted LocalTalk data can reduce the amplitude of the low frequency bits to a relatively small fraction of their initial amplitude, thus the distorted signal has less low frequency content. As shown in Fig. 6.17, the distorted LocalTalk signal C can be easily obtained by adding a smaller signal B which is a delayed inverse copy of A to the original signal A.

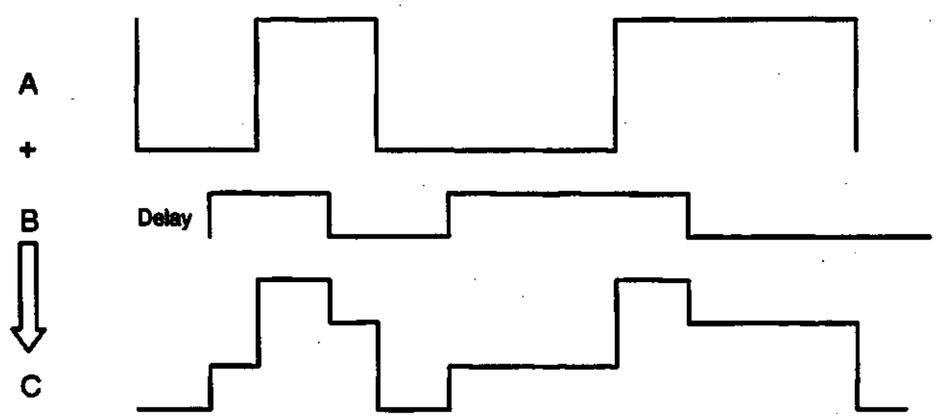


Fig. 6.17 Formation of a distorted signal

In order to show the improvement of the data transmission at long distance by using the predistorter, experiment systems shown in Fig. 6.18 are set up to obtain eye diagrams. The system shown in Fig. 6.18A is used to record the eye diagrams of the predistorted data. The unpredistored eye diagrams are recorded by the system shown in Fig. 6.18B. Both systems have the same high pass filter A and A1.

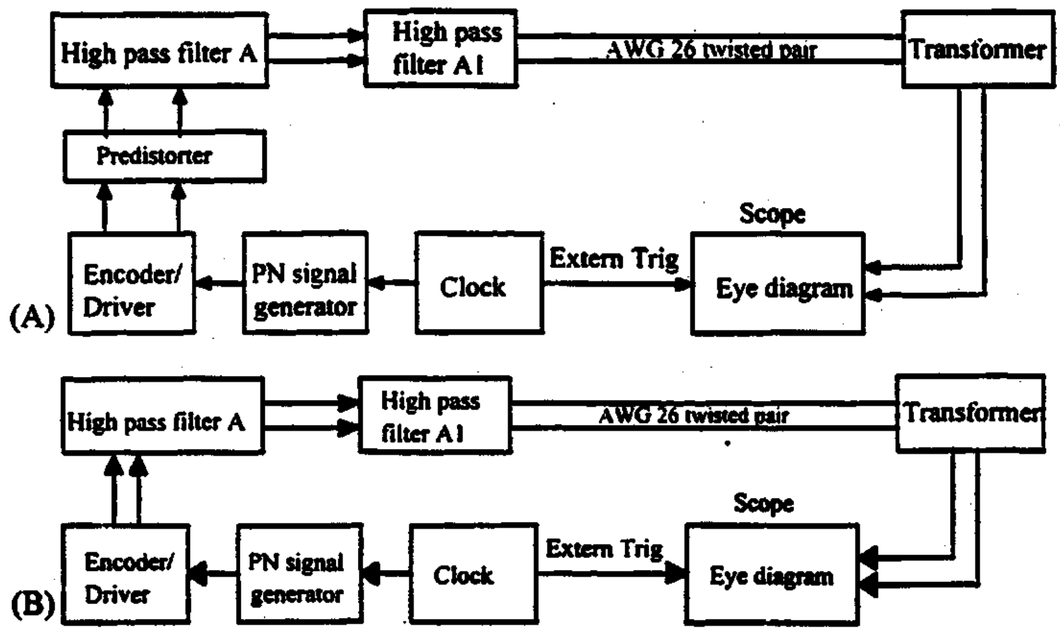


Fig. 6.18 The systems for obtaining eye diagrams with predistorter

The eye diagrams of the predistorted and the unpredistorted data are shown in Fig. 6.19. When the cable length is 3300 m, both the Eye 1 of the predistorted data and the Eye 2 of the unpredistorted data have open eyes, but the eye opening of the predistorted data is slightly larger than that of the unpredistorted data. When the cable length extends to 4000 m, the eyes of the unpredistorted data are closed (see Eye 4), but the predistorted data still has two open eyes (see Eye 3). The improved eye diagrams of the predistorted data can simplify the design of the equalizer at the receiver end.

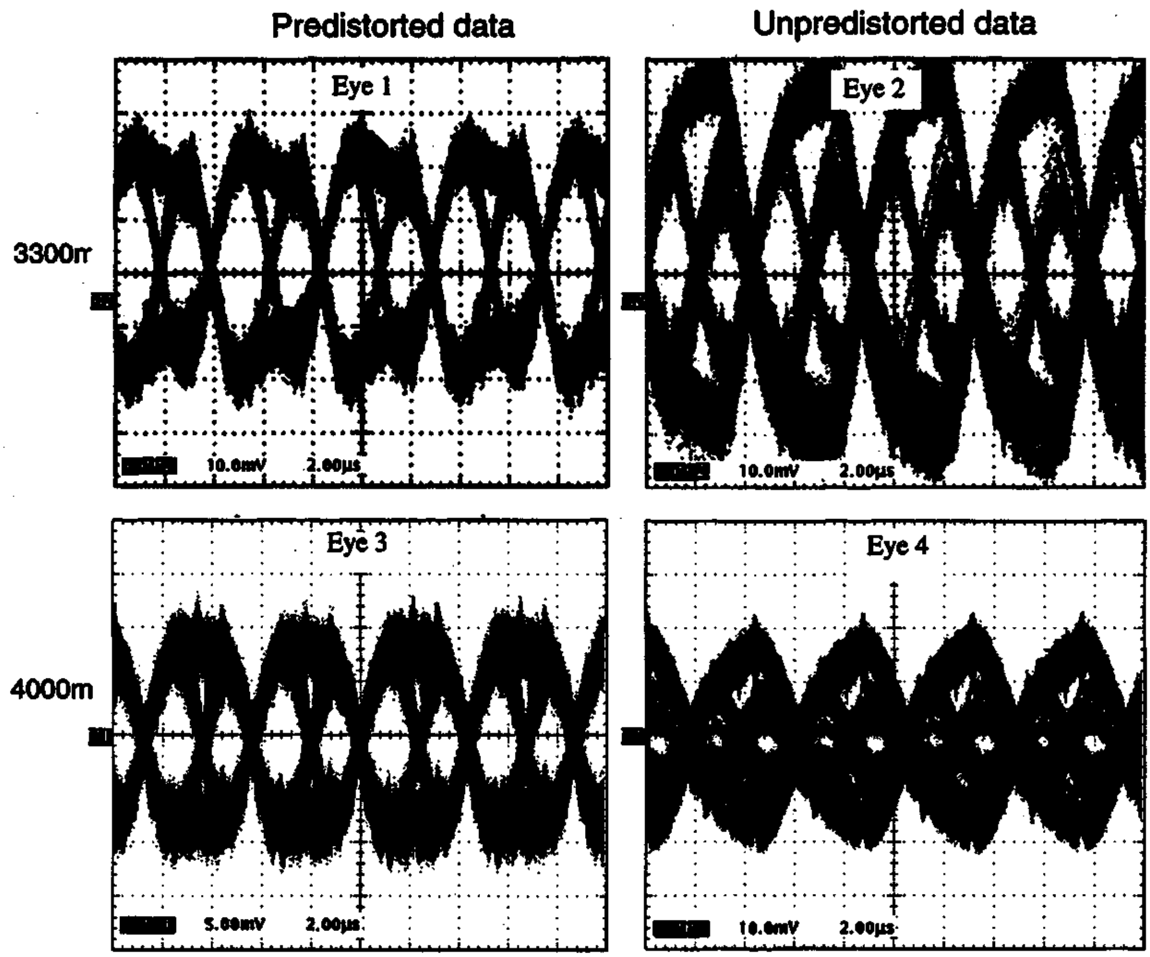


Fig. 6.19 Eye diagrams of data with predistortion and without predistortion

6.3.2 Circuits of the Predistorter and Equalizer

The circuits of the interface with the predistorter and equalizer are shown in Fig. 6.20. High pass filter A is the same as that in Fig 6.7. High pass filter A, A1 and the low pass filter B are still used to prevent the interference between the data and the voice. Because the output amplitude of the predistorted data is small, a central output grounded 1 to 1 ratio transformer is used here to achieve better circuit balance.

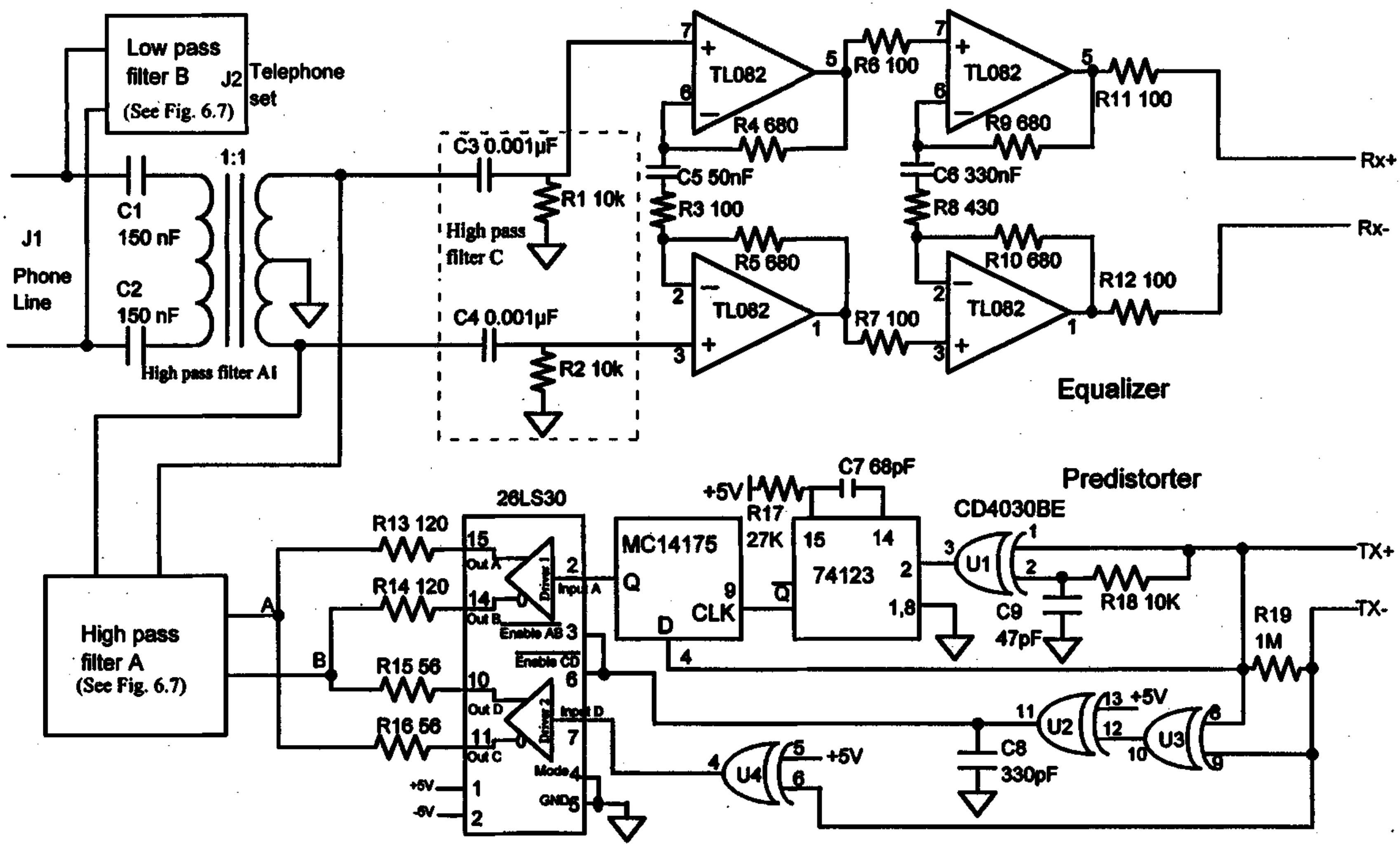


Fig. 6.20 New LocalTalk interface with predistorter and equalizer

The predistorter is combined with a delay circuit [32] and a sum circuit. MC14175, 74123 and the U1 in CD4030 chip compose the delay circuit. Balanced data is input from TX+ and TX-. When the data comes in, XOR gate U3 outputs '1' and U2 outputs '0' to enable the Driver 1 and 2 in 26LS30 chip. C8 is used to erase the glitch in the enable signal. U4 inverts the data in Tx- and is connected to the input of Driver 2. The XOR gate U1, C9 and R18 perform as a dual edge detector that can change the transitions of the TX+ data to narrow pulses (See Fig. 6.21B). After these pulses go through the multivibrator (74123), they are expanded to about 1.2 μs equal wide pulses, the width of the pulses are adjustable by C7 and R17 [33]. These pulses are output to the D Flip-Flop (MC14175) as the clock (See Fig. 6.21C). When the clock has a Low-to-High transition, it loads the TX+ data and output to the RS-422 driver (26LS30). Because each Low-to-High transition has a 1.2 μs delay to each original TX+ transition, therefore the delayed output data is the same as the original TX+ data (See Fig. 6.21 wave D).

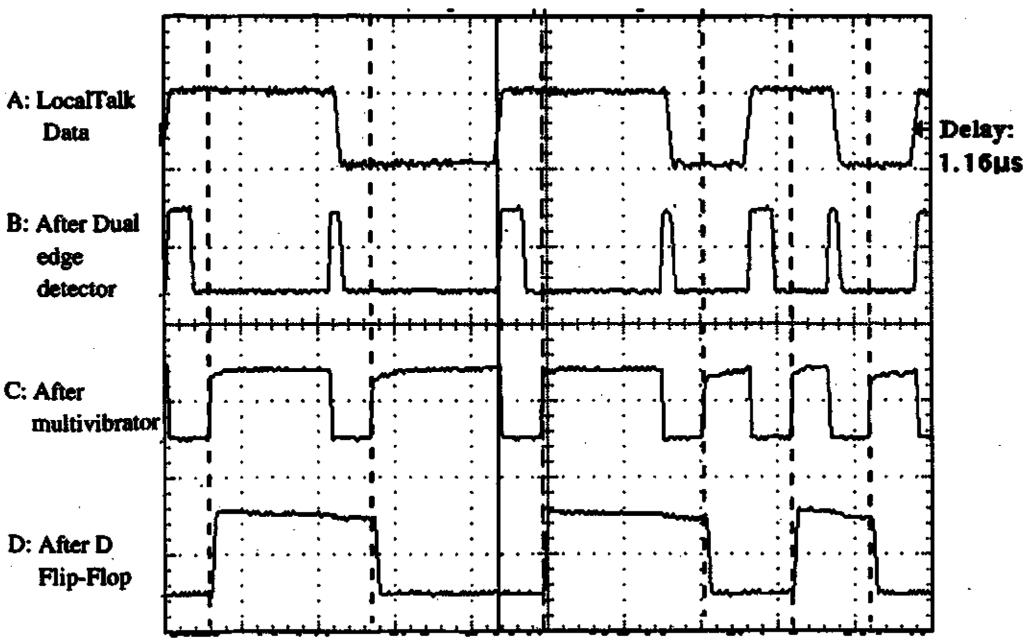


Fig. 6.21 Timing diagram of the delay circuit

The RS-422 line driver (26LS30) and R13~R16 form the sum circuit. Driver 2 drives the undelayed data to R15 and R16, while its delayed copy is output to R13 and R14. The positive and inverse outputs of Driver 1 are connected to the inverse and positive outputs of Driver 2. The output voltages (referenced to ground) at point A and B can be calculated as Eq. 6.2 (See Appendix 3 for detail explanation)

$$V_A = \frac{R_{16}}{R_{13}+R_{16}} V_{O1} + \frac{R_{13}}{R_{13}+R_{16}} (-V_{O2})$$
$$V_A = -V_B \qquad \text{Eq. 6.2}$$

where voltage V_{O1} is the positive output of Driver 1.
voltage V_{O2} is the positive output of Driver 2.

The output of the sum circuit (voltage between point A and B) is shown in Fig. 6.22.

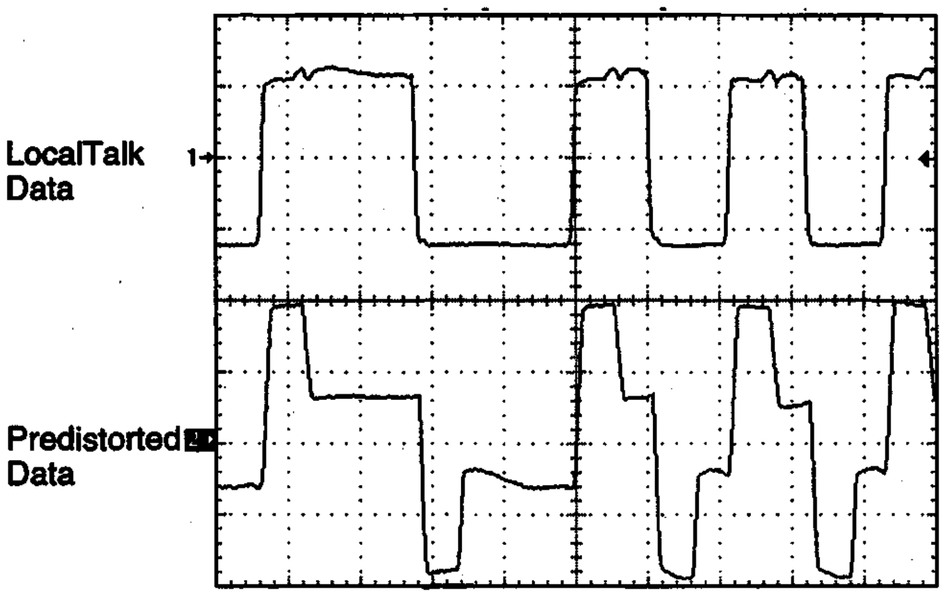


Fig. 6.22 Predistorted signal

The distorted signal still has noise on voice band. The high pass filter A and the low pass filter B are still needed to eliminate the voice band noise. These filters are the same as those in Fig. 6.7.

Because of the improvement caused by the predistorter, the eye diagrams of the received signal are fairly good after long distance transmission, but the amplitude of the received signal is very small. Therefore, a high gain equalizer is required at the receiver. As mentioned in Section 6.2.2, the bandwidth of the amplifier will decrease with the increase of the gain. In order to maintain the bandwidth and enough gain, a two-stage amplifier is used in the equalizer. With this two-stage equalizer, it is easy to have up to 35 dB gain at high frequency range. The Curve A in Fig. 6.23 shows the frequency response of the two-stage equalizer. Although the gain of the amplifier at high frequency is what we want, the gain at low frequency, especially in voice band, is not acceptable. Therefore, the high pass filter C is used to reduce the amplifier's gain at the low frequency range. The frequency response of the two-stage equalizer with the high pass filter C is shown in Fig. 6.23 Curve B. It is easy to see that the high pass filter C reduces the gain at the voice band remarkably.

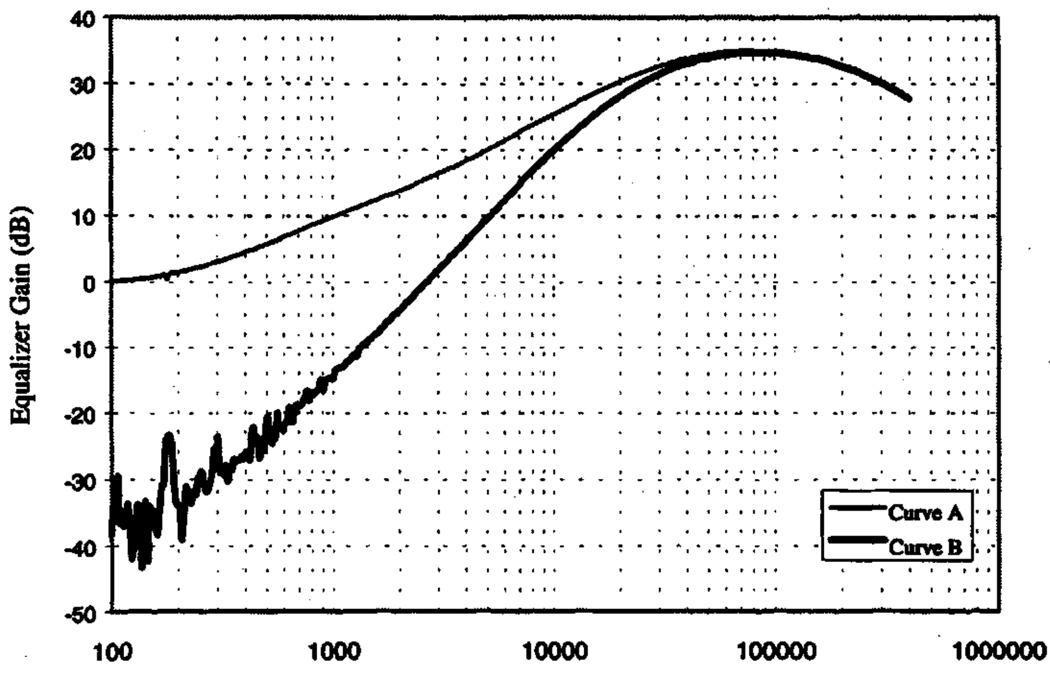


Fig. 6.23 Equalizer gain vs. frequency

6.3.3 Test Results

A system which is similar to Fig. 6.12 is used to record the eye diagrams, but the interface module used in Fig 6.12 is substituted by the interface module with both equalizer and predistorter. The eye diagrams of the received signal at 3300 m and 4000 m are shown in Fig. 6.24. At 4000 m, Eye 3 is the eye diagram of the received signal before equalizer and Eye 4 is the eye diagram of the same signal after the equalizer. Eye 3 and Eye 4 have similar shape, but Eye 4 has much larger eye opening. At 3300 m, Eye 1 and Eye 2 have similar relationship as that of Eye 3 and Eye 4.

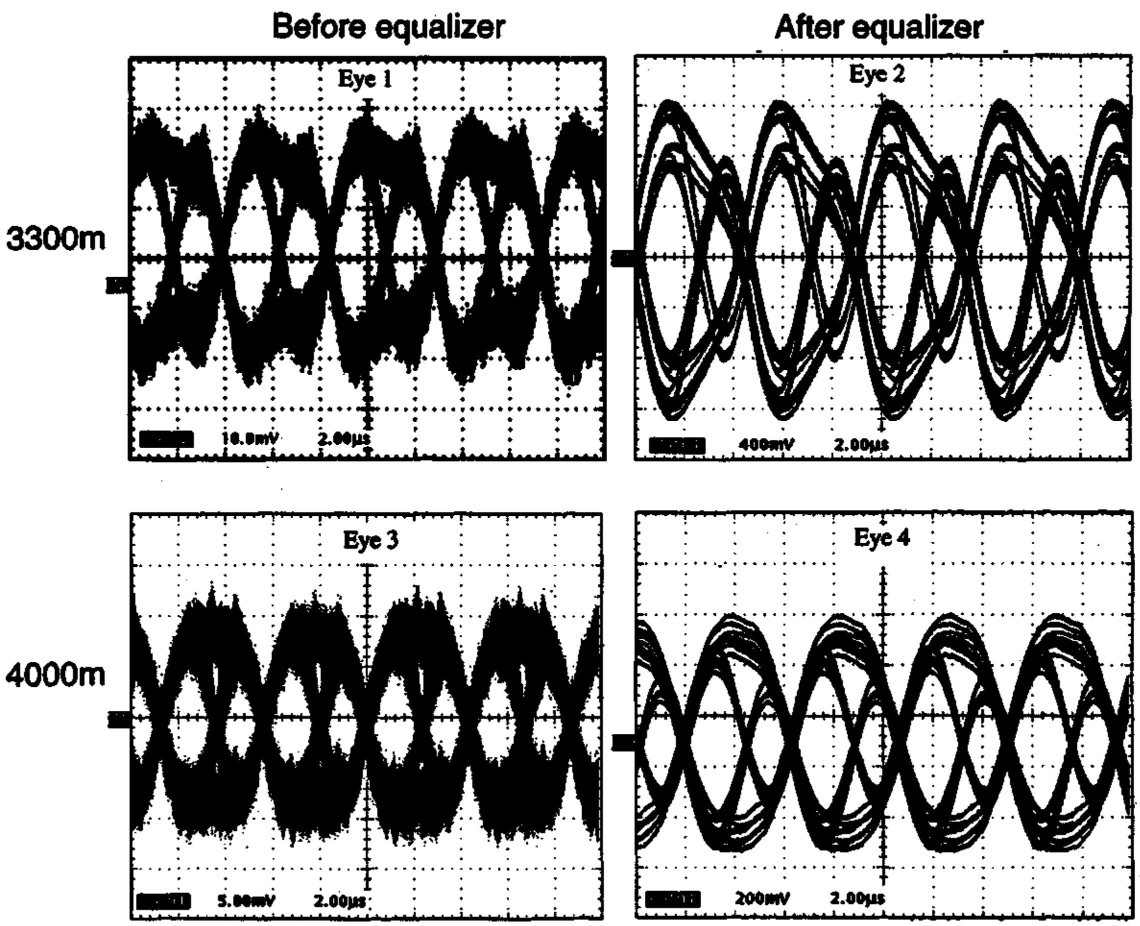


Fig. 6.24 Eye diagrams of received data at 3300 m and 4000 m

In order to test the performance of the circuit, similar test systems and test steps used in Section 6.2.3 are adopted here. The test results show that the predistorter and the equalizer interface module can work from 3300 m to 4000 m with 0.0% packets dropped and can download the 472 kbytes file in 41 seconds or so. When the data is transmitted through this interface module, data has even quiet noise in voice band. In the worst case, the SAGE 930A communication test set only can record 2 dBrnC noise that is generated by the data transmission. In order to test the effect of voice interference to the data transmission, a 1 kHz tone is transmitted simultaneously with the data. The interface module still can transmit and receive data without any packets dropped at 3300 m and 4000 m with the power of the tone increased up to 4 dBm.

6.4 Solutions to the Bridged Tap

Bridged taps are very common in telephone networks. Telephone companies often connect a pair of open circuit transmission lines to the subscriber loop between the central office and the terminal equipment. The open circuit pair is called bridged tap as shown in Fig. 6.25.

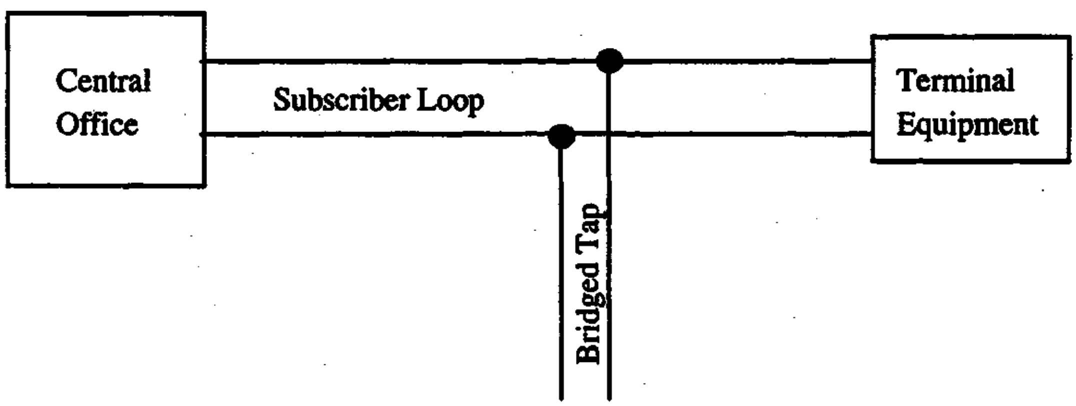


Fig. 6.25 Illustration of the bridged tap

In Chapter 3, the transmission theory shows that the open circuit transmission line reflects all incoming signals. When the length of the open circuit is about 1/4 of the wave length, the reflected signal will cause significant distortion to the signal transmitted in the subscriber loop.

The 1/4 wave length of the voice band signal is over 15 km. The bridged tap usually is not that long, and even if the bridged tap happens to be 15 km, the reflected signal is largely attenuated after the long round trip. However, the wave length of the LocalTalk data can be very short. The first null of the LocalTalk data spectrum is about 430 kHz, 1/4 wave length of this frequency is as short as 120 m. Since the main power of the LocalTalk data is concentrated from 30 kHz to 300 kHz, The bridged taps around 170 m or longer are of concern. Fig. 6.26 shows the frequency response of the 1500 m AWG26 cable with a 200 m bridged tap.

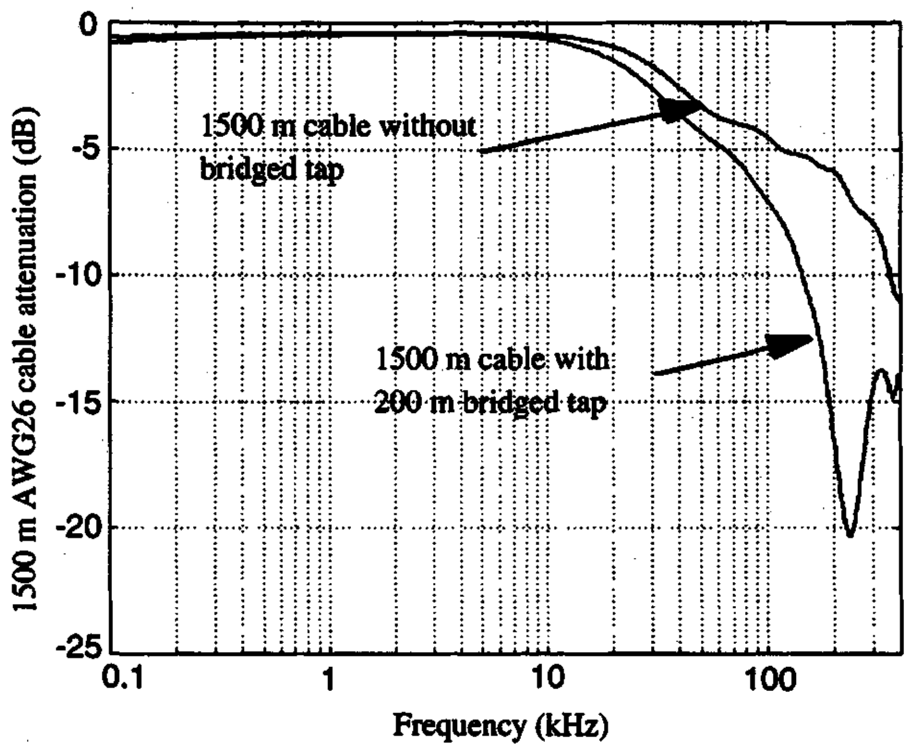


Fig. 6.26 1500 m cable attenuation with 200 m bridged tap

The frequency response of the cable with bridged tap has its first sharp attenuation peak around 250 kHz. This unexpected attenuation can significantly affect the eye diagram that is recorded at the terminal equipment side. Fig. 6.27 shows the effect of the bridged tap to the LocalTalk data. Eye A is the LocalTalk data being transmitted through the 1500 m cable and equalized without bridged tap, Eye B is the bridged tap version. Obviously the bridged tap is a problem to the data transmission.

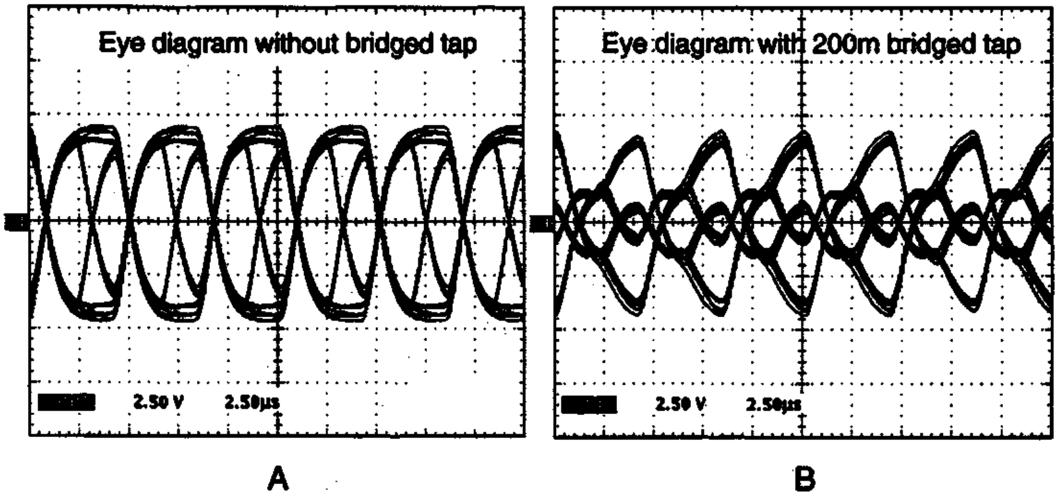


Fig. 6.27 Eye diagrams with/without bridged tap

According to the transmission line theory, the ideal solution to the bridged tap is to use a match terminator whose impedance is exactly the same as the characteristic impedance of the cable. However, this is not possible and not necessary. Some easier and effective methods are discussed here.

6.4.1 Solution 1: Pure Resistor Termination to the Bridged Tap

The Fig. 3.8 in Chapter 3 shows that the characteristic impedance of the phone cable tends to 100 Ω with increasing frequency. Thus, a 100 Ω resistor can be used as a terminator to eliminate or decrease the high frequency signal reflection. A 200 m bridged tap with a 100 Ω resistor termination is shown in Fig. 6.28.

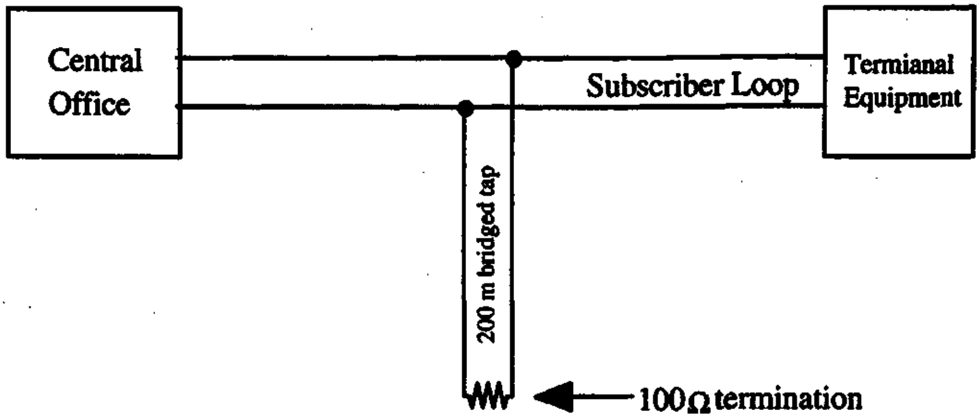


Fig. 6.28 Subscriber loop with 200 m pure resistor terminated bridged tap

Fig. 6.29 shows the frequency response of the 1500 m cable with a 200 m terminated bridged tap, and Fig. 6.30 shows the eye diagram, which is recorded at the terminal equipment side, of the transmitted signal after 1500 m cable with a 200 m terminated bridged tap.

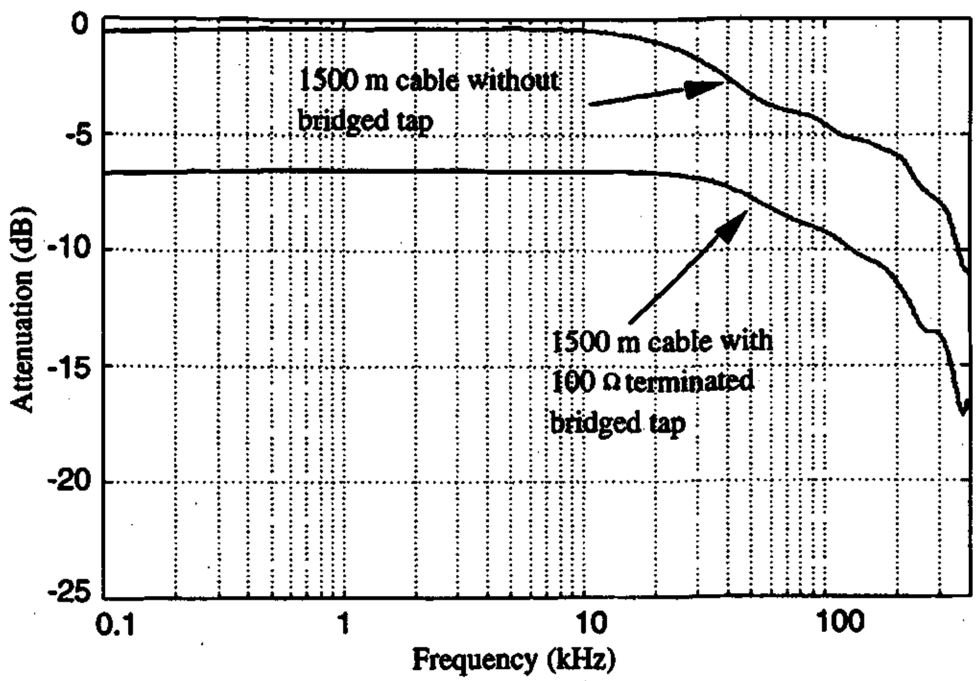


Fig. 6.29 The attenuation of 1500 m cable with 200 m bridged tap terminated by a 100 Ω resistor

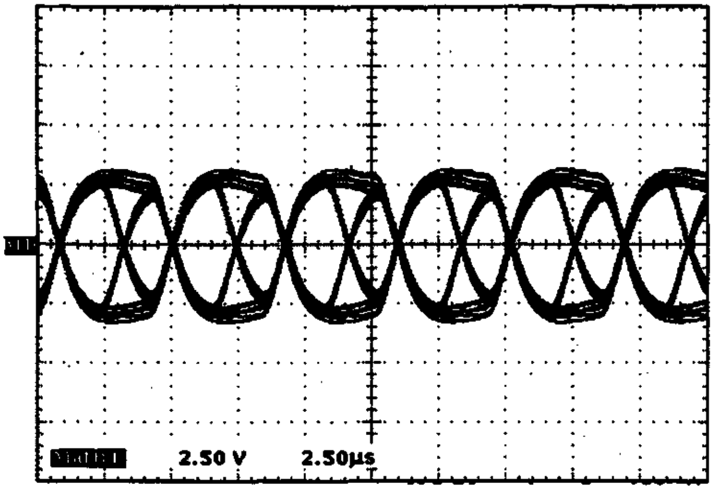


Fig. 6.30 Eye diagram of 1500 m cable with 200 m bridged tap terminated by a 100 Ω resistor

In Fig. 6.29, The terminated bridged tap doesn't have sharp attenuation to the cable and the eye diagrams are good in Fig. 6.30, but the problem with this method is that the termination causes much attenuation in the voice band and brings trouble to the telephone network too. The 100 Ω terminator can make the central office treat the terminal equipment on the subscriber loop as being off-hooked.

6.4.2 Solution 2: R-C Termination to the Bridged Tap

R-C termination is a better method to terminate the bridged tap. The termination is realized by a resistor in series with a capacitor (shown in Fig. 6.31).

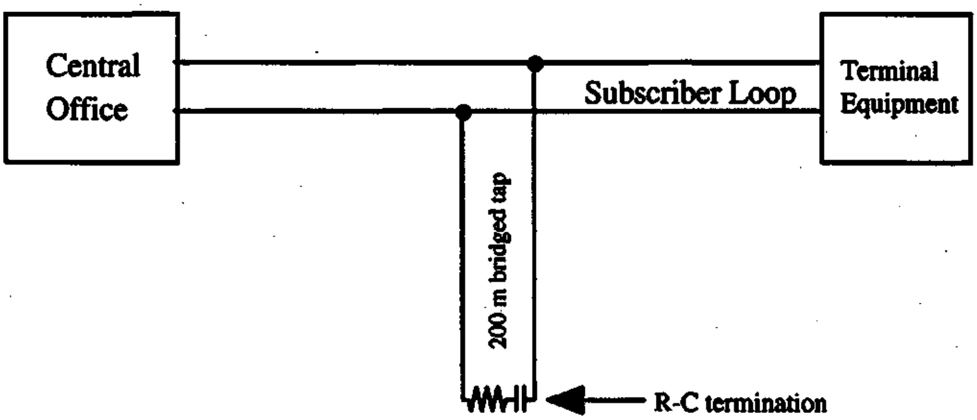


Fig. 6.31 Subscriber loop with 200 m R-C terminated bridged tap

Fig. 6. 32 shows the impedance of the R-C terminator whose R is 100 Ω and C is 0.022 uF. It is easy to see that the impedance of the R-C terminator is larger at low frequency band and tends to 100 Ω at high frequency. This characteristic can reduce the signal reflection of the LocalTalk data without disturbing the telephone network.

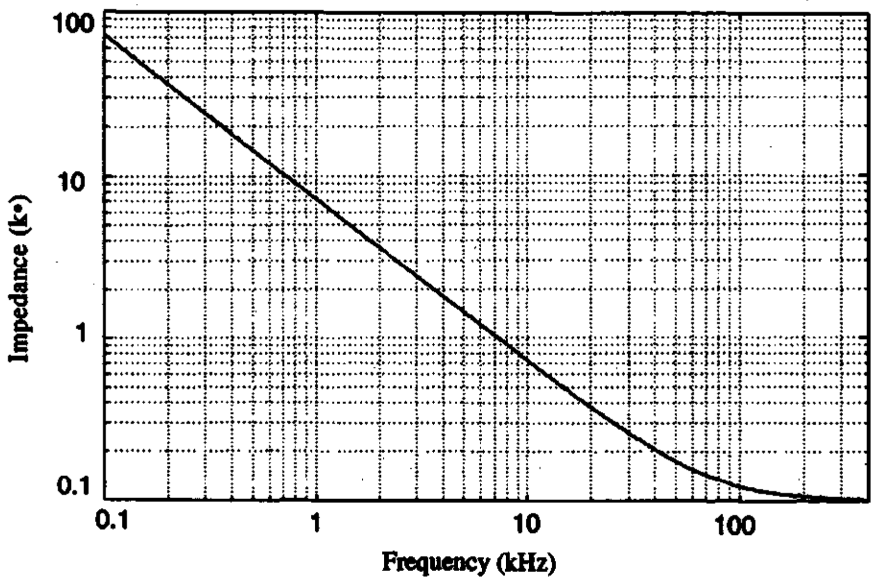


Fig. 6.32 Impedance of the R-C termination

The frequency response of the 1500 m cable with a 200 m R-C terminated bridged tap is shown in Fig. 6.33.

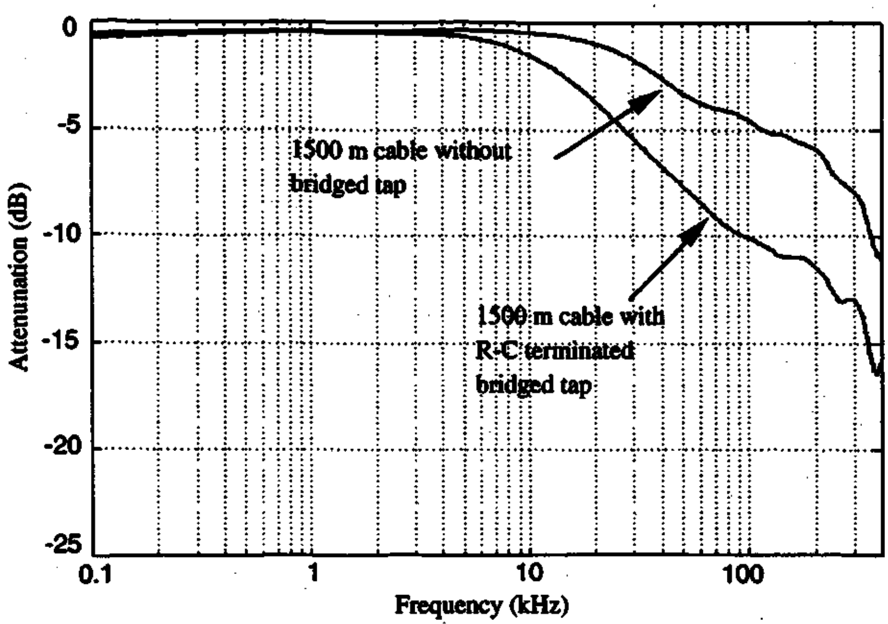


Fig. 6.33 Frequency response of a 1500 m cable with R-C terminated bridged tap

In Fig. 6.33, the R-C terminated bridged tap curve is almost identical to the response without bridged tap at voice band and there is no sharp attenuation peak at high frequency.

The eye diagram of the LocalTalk data transmitted after the 1500 m cable with a 200 m R-C terminated bridged tap is shown in Fig. 6.34. Again, this eye diagram is recorded at the terminal equipment side. The eye diagram shows that the R-C termination has almost the same shape as the 100 Ω terminator, and with this eye diagram, the receiver can receive data correctly.

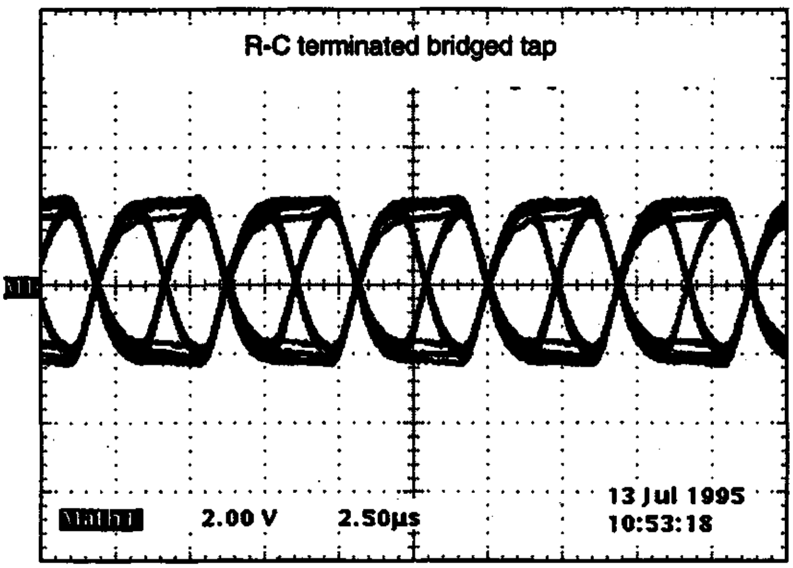


Fig. 6.34 Eye diagram of a 1500 m cable with 200 m R-C terminated bridged tap

6.5 Summary

In this chapter, We have presented two methods which can extend the LocalTalk communication distance and two solutions to the bridged taps.

The first method which can extend the range of the LocalTalk transmission is realized by an equalizer working with a new LocalTalk interface module that was introduced in Chapter 5. The equalizer is a frequency domain equalizer which has advantages of low cost and reliability. Because the LocalTalk uses a balanced signal, the balance of the equalizer circuit is an important issue in the equalizer design. In this project, LocalTalk data is transmitted simultaneously with the voice, thus the equalizer circuit is designed to obtain good equalization results to the data and avoid causing voice interference to the data. Test data shows that the equalizer has satisfactory performance at distances from 0 m to 3300 m.

The second method which is used to extend the data transmission to longer distances is conducted by a new LocalTalk interface module with a predistorter and a two-stage equalizer. In this system, the transmitted data is distorted by the predistorter and has better eye diagrams at longer distances. The two-stage equalizer can give received data signals sufficient compensation within a wider bandwidth. The new LocalTalk interface module with the predistorter and the equalizer can work properly at distances from 3300 m to 4000 m.

The pure resistor and the R-C termination are two solutions to the bridged taps which are longer than 150 m. Although both pure resistor termination and the R-C termination have similar performances to the bridged tap, the R-C termination is recommended for its trouble free to the telephone networks.

Chapter 7 Conclusion

This thesis presents a new high speed network access technology. This high speed network access method can transmit 230 kbits/s data simultaneously with voice on the existing telephone line at very low cost.

The LocalTalk local area network protocol, which is used in this project, is a well established robust protocol. In Chapter 2, the thesis has shown that the LocalTalk protocol can operate normally at range up to 9.6 km and the LocalTalk data is suitable for simultaneous transmission with voice. The use of the LocalTalk protocol also simplifies the design of this project and make it possible that the new network access method be mass deployed.

In Chapter 5, the thesis has shown that carefully designed high pass and low pass filters can be used to reduce the low frequency power of the data signal and prevent the voice and the subscriber loop signaling from interfering with the data. The high pass filters working with the low pass filter can reduce the LocalTalk data noise, which causes interference to the voice communication, from 81 dBrnC to 6 dBrnC and can also support normal telephone conversation without disturbing the data transmission. Experiments show that the new LocalTalk interface module with high pass and low pass filters can work very well at short distances.

In Chapter 6, the thesis has investigated methods of extending the range of the LocalTalk transmission. In order to achieve the data transmission distance that is longer than 2 km, an equalizer that can compensate for the distortion caused by the subscriber line is required at the receiver end. The equalizer shown in Fig. 6.7 is not complex and expensive, but it has satisfying performance at distances up to 3.3 km. For longer distance data transmission, in addition to a equalizer being used at the receiver, a predistorter is needed to predistort the data signal before it is transmitted. With predistortion and equalization, the data transmission distance can be increased up to 4 km.

The impairment caused by bridged taps has also been investigated in Chapter 6. The bridged tap in an analog telephone network is not harmful, but it is a disaster in a high speed data network. In this project, the experimental results show that the bridged tap should be shorter than 150 m, otherwise it must be terminated for signal frequencies of data transmission.

Future improvements of this project can be achieved by updating the receiver circuit in the interface module. The high pass filter C in Fig. 6.7 and Fig. 6.20 can be substituted by a more powerful band pass filer. The pass band of the band pass filter should be 20 kHz to 400 kHz with sharp transition curves to stop band. The purpose of the pass band filter is to significantly reduce the interference coming from the voice at the low frequency band and the radio interference at the high frequency band. Experiments indicate that, different models of Macintosh computer and PC LocalTalk boards have different performances on the same equalized signal. In order to ensure consistent good performances under the same line condition for different LocalTalk receivers, an extra circuit is needed to modify the received equalized signal to a standard signal level which the LocalTalk receiver in every Macintosh or PC LocalTalk board can decode correctly.

TRLabs working with SaskTel have conducted some field trials on this new network access technology. The results have been satisfactory and indicate strong commercial potential.

References

1. *Access 230 Technology Transfer Opportunity*, TRLabs, 1996.
2. Charles K. Summers, *ISDN Implementor's Guide -- Standards, Protocols, and Services,* McGraw-Hill, NY, 1995.
3. B. R. Saltzberg, T, R, Hsing, *Guest Editorial*, IEEE Journal on Selected Areas in Communications, Vol. 9. No. 6, August 1991.
4. William C. Barlet, *A Copper-plated Full Service Network*, Telephony, January 15, 1996.
5. J. W. Lechleider, *High Bit Rate Digital Subscriber Lines: A Review of HDSL Progress*, IEEE Journal on Selected Areas in Communications, Vol. 9, No. 6, August, 1991.
6. David E. Dodds, *Internet Access Study*, TRLabs Saskatoon Report, June 7,1996.
7. *Cable Modem Frequently Asked Question*, http://www.cox.com/modemfaq.html
8. Ray Smith, *Catching up*, Telephony, June 24, 1996
9. David E. Dodds, Greg Erker and Graham Bradley, *Internet Access at 230 kb/s - Hardware Trial*, Paper in Preparation, 1997.
10. *Farallon™ PhoneNET® Connector User's Guide*, Farallon Computing, Inc. Berkeley, CA.
11. G. S. Sidhu, R. F. Andrews, and A. B. Oppenheimer, *Inside AppleTalk*, 1st Edition, Apple Computer, Inc. 1986.
12. G. S. Sidhu, R. F. Andrews, and A. B. Oppenheimer, *Inside AppleTalk*, 2nd Edition, Apple Computer, Inc. 1990.

13. A. S. Tanenbaum, *Computer Networks*, 2nd Edition, Prentice Hall, Englewood Cliffs, New Jersey, 1988.

14. *Macintosh Family Hardware Reference*, Addison-Wesley Publishing Company, Inc.

15. B. M. Wirth, *Fiber Optic Extension of Apple LocalTalk Networks*, M.Sc. Thesis, University of Saskatchewan, Saskatoon, Saskatchewan, 1989.

16. Data Sheet, "AM26LS30", *Bipolar Microprocessor Logic and Interface Data Book*, Advanced Micro Devices, Sunnyvale, CA, 1985.

17. Data Sheet, "AM26LS32", *Bipolar Microprocessor Logic and Interface Data Book*, Advanced Micro Devices, Sunnyvale, CA, 1985.

18. Data Sheet, "Z8030/Z8530 ", *Zilog Serial Communication Controllers Data book*, Zilog, Campell, CA, 1994.

19. Leon W. Couch II, *Digital and Analog Communication systems*, Fourth Edition, Maxwell Macmillan Canada, Toronto, 1993.

20. Bell Telephone Laboratories, *Transmission Systems for Communications*, Vol. 1, Bell Telephone Laboratories, New York, N.Y.

21. David. E. Dodds, *EE 816 Telephony Notes*, University of Saskatchewan, Saskatoon, Saskatchewan, 1996

22. H. H. Skilling, *Electric Transmission Lines: Distributed Constants, Theory and Applications*, McGraw-Hill, New York, 1951.

23. AT&T, Bell Telephone Companies, and Bell Telephone Laboratories, *Telecommunications Transmission Engineering*, Vol. 2, 1st Edition.

24. AT&T, *Principles of Electricity Applied to Telephone and Telegraph Work*, AT&T, 1961.

25. J. G. Proakis, *Digital Communications*, 2nd Edition, McGraw-Hill, New York, 1991.

26. Shahid Qureshi, *Adaptive Equalization*, IEEE Communications Magazine, March, 1982.

27. W. D. Reeve, *Subscriber Loop Signaling and Transmission Handbook: Analog*, IEEE Press, New York, 1992.

28. *SAGE Instruments Model 930A Communications Test Set Operations Manual*, SAGE Instruments, Freedom CA, 1990.

29. Walter G. Jung, *IC Op-Amp Cookbook*, 3rd Edition, SAMS, Camel, Indiana.

30. Data sheet, *"TL082", Linear Circuits Data Book*, Vol. 1, Texas Instruments, Dallas, Texas, 1992.

31. A. X. Widmer, N. Y. Katonah, *United States Patent 3,980,826*, International Business Machines, Armonk, N.Y., 1975.

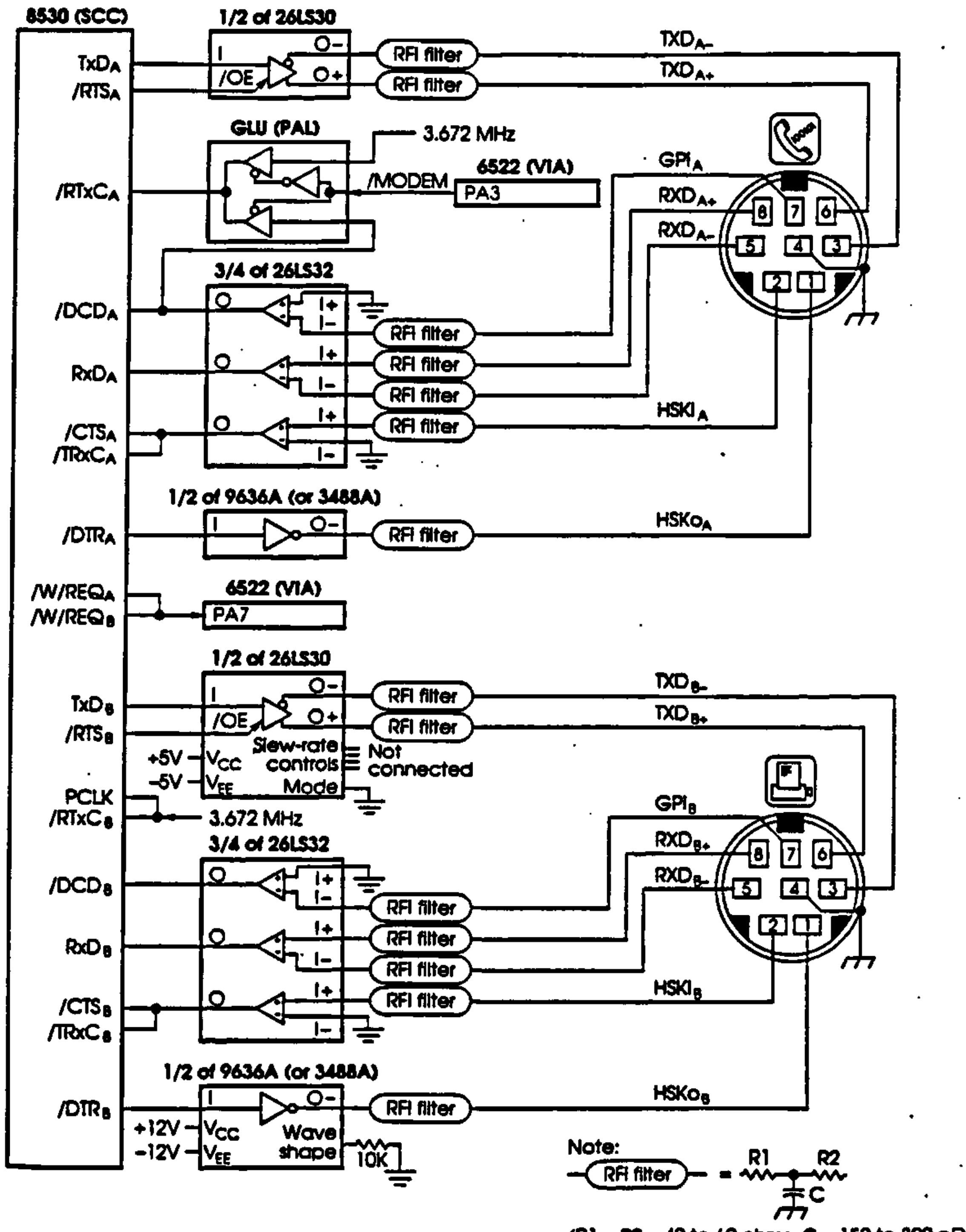
32. Rudolf F. Graf, *The Encyclopedia of Electronic Circuits*, 1st Edition, Vol. 3, Tab Books, Blue Ridge Summit, PA, 1985.

33. Data sheet, " SN74123", *TTL Logic Data Book*, Texas Instruments, Dallas, Texas, 1988.

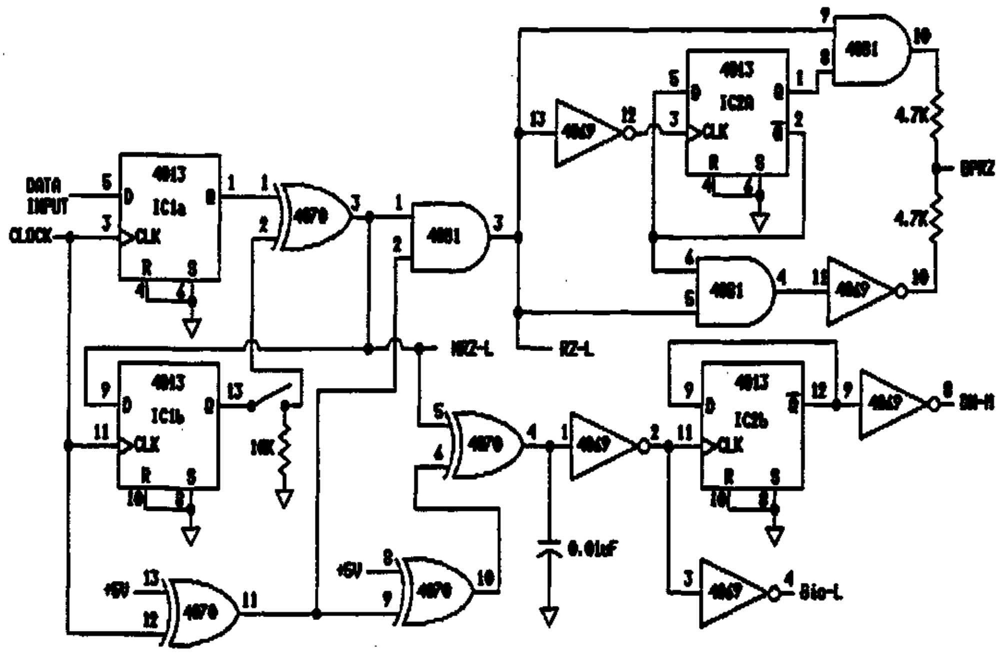
34. Whiteham D. Reeve, *Subscriber Loop Signaling and Transmission Handbook: Digital*, New York: IEEE Press, 1995.

35. Data sheet, "OPA2111", *Burr-Brown IC Data Book – Linear Products*, Burr-Brown Corporation, Tucson, AZ, 1994

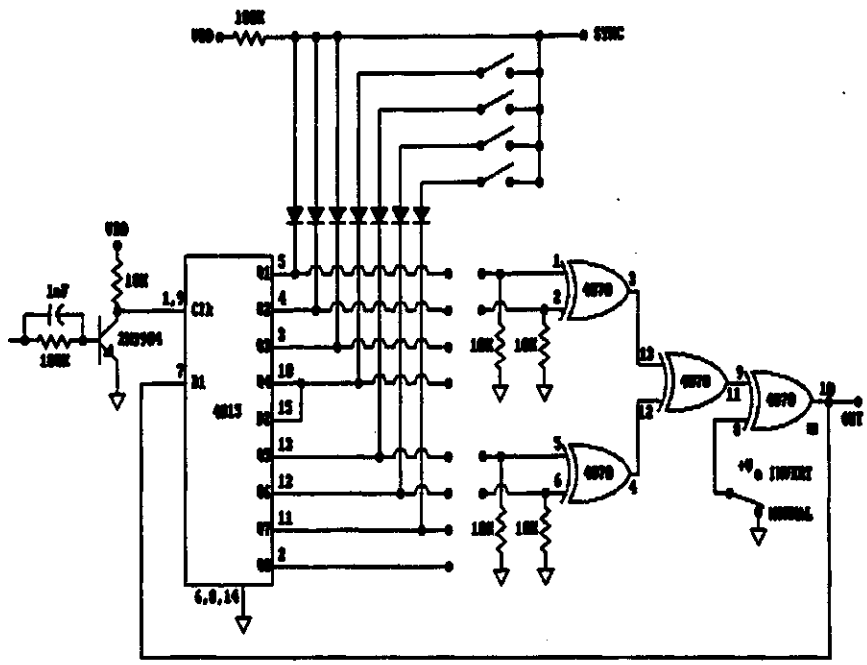
Appendix 1 Macintosh Serial Port



Appendix 2 Circuits of the Pseudo-Random Data Generator and the Baseband Encoder



Baseband Encoder



Pseudo-Random Data Generator

Appendix 3 Output Voltage of the Predistorter

The sum circuit of the predistorter is redrawn as Fig. A3.1.

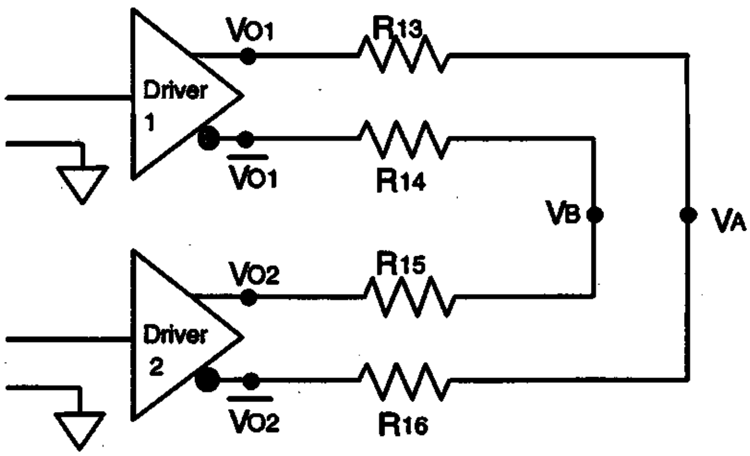


Fig. A3.1 The sum circuit of the predistorter

The outputs of Driver 1 satisfy

$$\overline{V_{O1}} = -V_{O1} \qquad \text{Eq. A3.1}$$

and the outputs of Driver 2 have

$$\overline{V_{O2}} = -V_{O2} \qquad \text{Eq. A3.2}$$

V_B and V_A can be calculated as

$$\begin{aligned} V_A &= V_{O1} - \frac{V_{O1} - \overline{V_{O2}}}{R_{13} + R_{16}} R_{13} \\ &= \frac{V_{O1} R_{16} + \overline{V_{O2}} R_{13}}{R_{13} + R_{16}} \end{aligned} \qquad \text{Eq. A3.3}$$

$$V_B = V_{O2} - \frac{V_{O2} - \overline{V_{O1}}}{R_{14} + R_{15}} R_{15}$$
$$= \frac{V_{O2}R_{14} + \overline{V_{O1}}R_{15}}{R_{14} + R_{15}} \qquad \text{Eq. A3.4}$$

Substitute Eq. A3.3 with Eq. A3.1 and Eq. A3.4 with Eq. A3.2, We get

$$V_A = \frac{V_{O1}R_{16} - V_{O2}R_{13}}{R_{13} + R_{16}}$$
$$= \frac{R_{16}}{R_{13} + R_{16}} V_{O1} + \frac{R_{13}}{R_{13} + R_{16}} (-V_{O2}) \qquad \text{Eq. A3.5}$$

and

$$V_B = \frac{V_{O2}R_{14} - V_{O1}R_{15}}{R_{15} + R_{14}}$$
$$= \frac{R_{14}}{R_{15} + R_{14}} V_{O2} + \frac{R_{15}}{R_{15} + R_{14}} (-V_{O1}) \qquad \text{Eq. A3.6}$$

According to the sum circuit of the predistorter,

$$R_{15} = R_{16} \qquad \text{Eq. A3.7}$$

and

$$R_{14} = R_{13} \qquad \text{Eq. A3.8}$$

Substitute Eq. A3.6 with Eq. A3.7 and Eq. A3.8, we get

$$V_A = \frac{R_{15}}{R_{14} + R_{15}} V_{O1} + \frac{R_{14}}{R_{14} + R_{15}} (-V_{O2}) \qquad \text{Eq. A3.9}$$

Comparing Eq. A3.6 to Eq. A3.9, we have

$$V_A = -V_B \qquad \text{Eq. A3.10}$$