



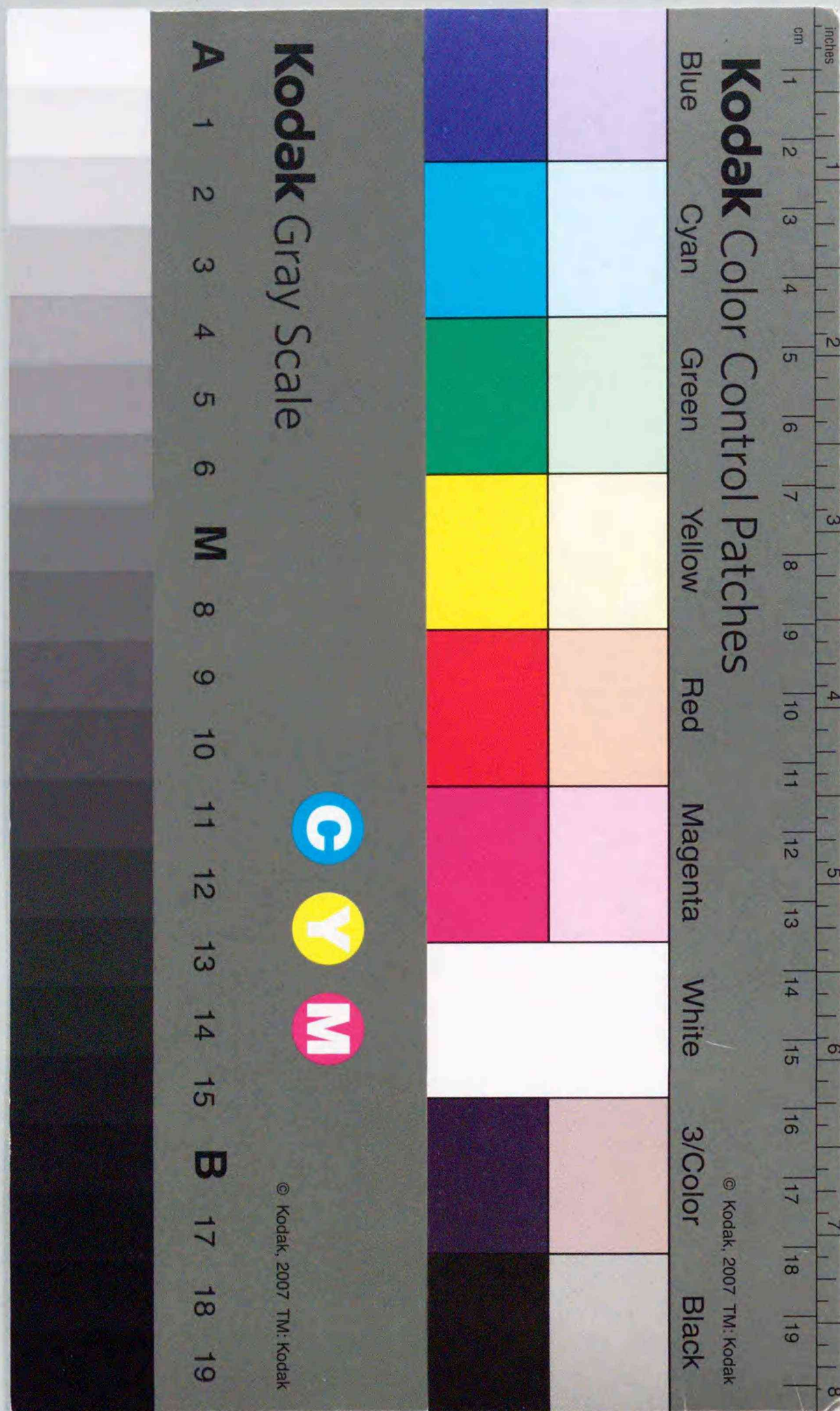
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ENHANCEMENT OF ULTRA-HIGH SPEED
TRANSMISSION TERMINAL EQUIPMENT

Yoshiaki Yamabayashi



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Chapter I

Introduction

1.1. Background

Telecommunication technology is being revolutionized by two of the greatest inventions of the 20th century, optical fiber and the integrated circuit (IC). Based on these inventions, large-capacity trunk transmission systems have been developed that offer reduced charges for long-distance and international calls. They are now becoming an important infrastructure for various telecommunication services such as electronic mail, Internet, video distribution, and many other newcomers.

Historically, the medium for propagating light was developed in two parts, first the waveguide structure and second the materials. In 1964, Nishizawa and Sasaki of Tohoku University proposed and patented a waveguide structure whose refractive index was maximum at the center of the core and decreased away from the center [1]. The material studies for optical-loss factors in glass by Kao and

Hockham led to the conclusion that a low-loss fiber could be produced by thoroughly removing all transition metals [2]. From the beginning, several advantages of the new medium over coaxial metal cable were obvious, such as lighter weight, smaller bending radius, and freedom from electric induction. However, it was still unclear at that time what wavelength would prove to be the most suitable for large capacity long-distance transmission. Following the above proposals, a lot of efforts have been made to exploit the untapped virtues of fiber, namely low-loss and broad-band characteristics. In reducing the optical loss, a milestone was placed by Kapron in 1970 with the value of -20 dB/km [3]. The race in manufacturing technology has almost been settled with two reports, one for -0.2 dB/km by Miya et al. in 1979 [4] and the other for -0.15 dB/km by Yokota et al. in 1985 [5]. Single-mode fiber, which is free from the band limitation, experienced by multimode fiber (modal dispersion), due to group-velocity difference among the modes was developed. The fiber has a 10 μm diameter core with 125 μm cladding, and exhibits the important fact that 1.3 μm light offers the fastest propagation speed. The characteristic wavelength is called the "zero-dispersion wavelength" of the medium, because the first derivative of the group-velocity, or wavelength dispersion, is null at that wavelength. In other words, pulse broadening due to the spectral width of the launched light is minimal at that wavelength. Furthermore, the 1.3 μm band was the only low attenuation window available at that time, because the ultimate low-loss window at 1.5 μm was blocked by the absorption caused by residual OH-bonds. Therefore, many transmission systems are being designed on the 1.3 μm window even now. Recent advances in both manufacturing and design have made it possible to shift the zero-dispersion wavelength to the 1.5 μm band, with an attendant increase in repeater spacing [6].

In step with fiber technology, optical device technology has also been developed. The first epoch in the laser field was the successful demonstration of continuous semiconductor laser oscillation at room-temperature by Hayashi et al.

in 1970 [7]. The semiconductor laser, or laser diode (LD), was originally the only practical way of realizing compact optical transmitters. The only commercially available solid-state light detector then was the Si-APD (Si-Avalanche Photodiode). GaAlAs system LDs emitting 0.85 μm light were initially developed to match the known detector response. They were followed by InGaAsP LDs to satisfy the requirement for a longer wavelength window, from 1.3 μm to 1.5 μm , which was more suitable for long distance transmission. A breakthrough was in achieving single-lateral mode oscillation, as this was necessary for efficient and stable coupling with single-mode fibers [8]. A narrow (μm order) stripe of the active region was etched and buried with inactive materials to form a single-lateral mode cavity. The next achievement was single-longitudinal mode oscillation by introducing a distributed-feedback (DFB) structure [9]. Instead of facing cleaved crystal edges to form a Fabry-Perot cavity, a region with corrugated refractive index or optical gain, the period of which equaled the optical wavelength along the propagation direction, was formed to make the cavity mode-selective. For long-wavelength window application, Ge-APD and then InGaAs-APD were developed, the latter was featured in terms of dark current and excess noise [10]. Recently, direct optical amplifiers based on Er-doped silica fiber as the gain medium have been developed to obtain broad-band and high-gain signal amplification beyond the constraint inherent in electrical amplification. The linear repeater composed of such an optical amplifier has made possible low cost and bit rate insensitive transmission systems [11].

Generally speaking, both larger transmission capacity and longer repeater spacing contribute to reducing communication cost. A larger capacity allows more users to share the same transmission line including intermediate and terminal repeaters and cables. Longer repeater spacing means fewer repeaters are needed to construct the line so that both the building and maintenance cost can be reduced accordingly. These benefits are the results of not only the low attenuation and excellent dispersion property of the transmission medium, but also of

the development of advanced circuit techniques for optical modulation/demodulation, amplification techniques (either optical or electrical), and digital high-speed processing circuits for transmission coding and decoding.

The first optical transmission systems introduced into Japan in 1981 were the F-32M and F-100M. The numerical value in each name (32 and 100) roughly indicates the information bit rate (precisely 32.06 Mbit/s and 97.72 Mbit/s, respectively). The multimode fiber structure employed was a graded-index core with the operational wavelength of either 0.85 μm or 1.3 μm . The former system utilized GaAlAs-LD as the light source and Si-APD as the receiver for 10 km maximum repeater spacing; the latter used InGaAs-LD and Ge-APD to realize 15 km spacing. These repeater spacing values well suited their main target of trunk lines in urban and suburban areas.

For trunk lines connecting large cities, the coaxial cable system DC-400M had been introduced in 1976 with the capacity of 397.2 Mbit/s. Its intermediate repeater spacing was 1.5 km. The optical version, F-400M, introduced in 1983, had 20 km spacing due to adopting single-mode fiber instead of coaxial cable. The 5760 telephone channels, originally transmitted using bipolar pulses in the electric systems, were now carried by unipolar return-to-zero (RZ) pulses, however the same signal frame, or multiplexing architecture with stuffing, was used in the optical version.

Both the capacity and spacing were upgraded thereafter to the first gigabit system, F-1.6G, in 1987, which utilized the combination of DFB-LDs and InGaAs-APDs. At the end of 1988, by incorporating opto-electronics and network synchronization technologies, ITU (International Telecommunication Union) agreed on the international standard interface named SDH-NNI, Synchronous Digital Hierarchy-Network Node Interface [12]. Three systems, FTM-156M, FTM-600M, and FTM-2.4G, were developed that complied with this specification. Two wavelength bands were utilized, 1.3 μm over conventional single-mode fibers for 40 km repeater spacing, and 1.5 μm over dispersion-

shifted fibers for the doubled spacing of 80 km. As of 1996, a system capable of 10 Gbit/s over 80 km of linear repeater and 320 km regenerator spacing is available. The code utilized is a scrambled non-return-to-zero (NRZ) unipolar pulse for an STM-64 frame which has been proposed as a natural extension of the SDH standard.

The coherent transmission technology was once studied intensively to increase the repeater spacing. To enable coherent detection, which utilizes the beat generation between two laser sources at the transmitter and receiver, direct measurement of the temporal coherent characteristics of semiconductor lasers is important. A series of interferometric measurements revealed relaxation oscillation spectra being associated with the multilongitudinal laser spectra with just DC feeding [13]. An extended study was made on utilizing coherent modulation in frequency-modulated semiconductor lasers [14]. This letter, ref. 14, reported the application of measuring the polarization dispersion of polarization preserving fibers. Polarization dispersion can be determined from the modulation conditions that give depolarization at the fiber end, when the fast and slow axis components are launched simultaneously at the input port.

Higher bit-rate transmission systems with a capacity of several tens of gigabit per second would require ultrafast waveform measurements of light pulses with a resolution of ten picoseconds or less. Conventional waveform measurement system using a fast photodetector followed by an electrical sampling circuit can not achieve such resolution because of their bandwidth limitation. Streak cameras can yield about 5 ps resolution, however, they are limited to applications that have specific repetition frequencies demanded by the measurement system. This conditional usage prevents them from being applied to the "eye-pattern" evaluation of transmission light pulses. The optical sampling technique utilizing a combination of sum-frequency generation in a nonlinear crystal and ultrafast laser pulses offers an attractive method that avoids the problems associated with high-speed electronics [15]. Laser diodes are superior to

other lasers such as dye lasers and solid-state lasers, in terms of not only size but also external tunability of the pulse generation frequency. Gain-switched, pulse-compressed distributed-feedback-laser diodes (DFB-LDs) have been used to achieve a time resolution of 8.7 ps; a bulk nonlinear crystal of LiIO_3 was used [16]. The usage of a LiNbO_3 waveguide enhanced the conversion efficiency by a factor of 3, and realized continuous matching between the fundamental and sum-frequency mode. The latter benefit originates from the fact that the sum-frequency light is generated in the Cherenkov radiation mode [17]. These works, as well as signal-to-noise ratio analyses, are summarized in a paper [18]. Active mode-locked fiber lasers have, as another compact pulse source, been studied to generate transform-limited picosecond pulses [19]. Optical amplification and newly developed high-efficiency crystals have made the averaging process for SNR (Signal to Noise Ratio) improvement unnecessary. In other words, the eye-diagram plotting or the display of each sampled point in a way similar to that done by conventional sampling oscilloscopes is now possible [20].

1.2. Multiplexing Methods for Transmission

Multiplexing methods can be categorized into two major categories, analog and digital. The digital hierarchy is divided further into plesiochronous and synchronous schemes. The former developed for asynchronous networks, remained in service even after synchronization of the lower stages of the hierarchy. The SDH standard rendered plesiochronous schemes obsolete by synchronizing the whole network up to the highest stage of the hierarchy.

1.3 Analog Multiplexing

Prior to the advent of digital technology, analog multiplexing was the only choice available for economical telephone networks. The 4 kHz voice channels are lined up on the frequency axis, while avoiding spectral overlap. As one

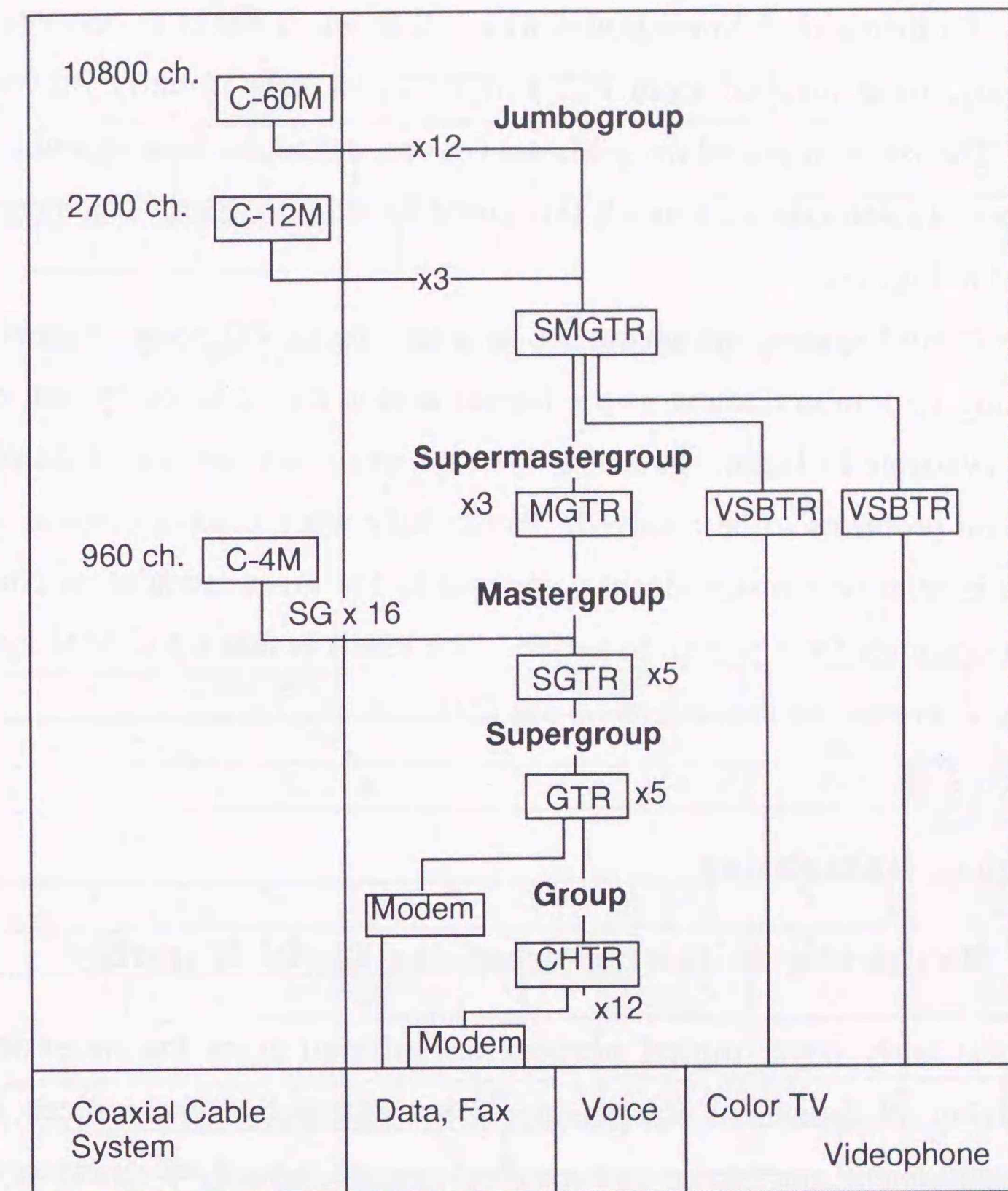


Fig. 1-1 FDM analog network hierarchy

example, the voice signal first premodulated a 130 kHz carrier, and then 12 of them were spaced between 60 kHz and 108 kHz, forming a group. Five groups, 60 channels in total, called the Supergroup, were distributed from 312 kHz to 552 kHz,. Furthermore, 5 Supergroups with 8 kHz guard bands in between, the Mastergroup, were mapped from 812 kHz to 2,044 kHz to carry 300 voice channels. The last step placed three Master Groups, called the Jumbogroup, from 8,516 kHz to 12,388 kHz with an 88 kHz guard band in between. The hierarchy is depicted in Fig. 1-1.

The C-60M system, which carries as many as 10,800 voice channels by multiplexing 12 Jumbo Groups, is the largest analog transmission system commercially available in Japan. In analog systems, cross talk and signal distortion are the main problems to be resolved. Particularly when coaxial cable is used, cross talk is relatively insignificant compared to the cross-modulation components that originate from signal distortion. The result is that the C-60M system has a maximum repeater spacing of 1.5 km [21].

1.4. Digital Multiplexing

1.4.1. Multiplexing in the Plesiochronous Digital Hierarchy

In the early days, digital networking suffered from the necessity of synchronizing all dependent equipment. If not externally synchronized, each piece of equipment must have its own clock source, usually a quartz crystal. Since each crystal guarantees only the frequency range within which it oscillates, the clocks have slightly different frequencies. As a result, data pulses can not have exactly the same period from channel to channel, as shown in Fig. 1-2. In digital time division multiplexing (TDM), however, clock frequencies must be unified. Successful reclocking can be done if the unified clock frequency is

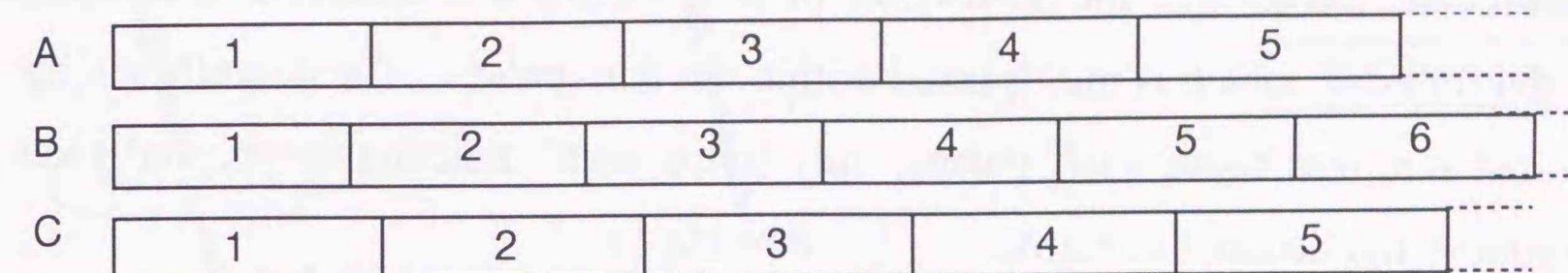


Fig. 1-2 Signal pulses from unsynchronized equipment

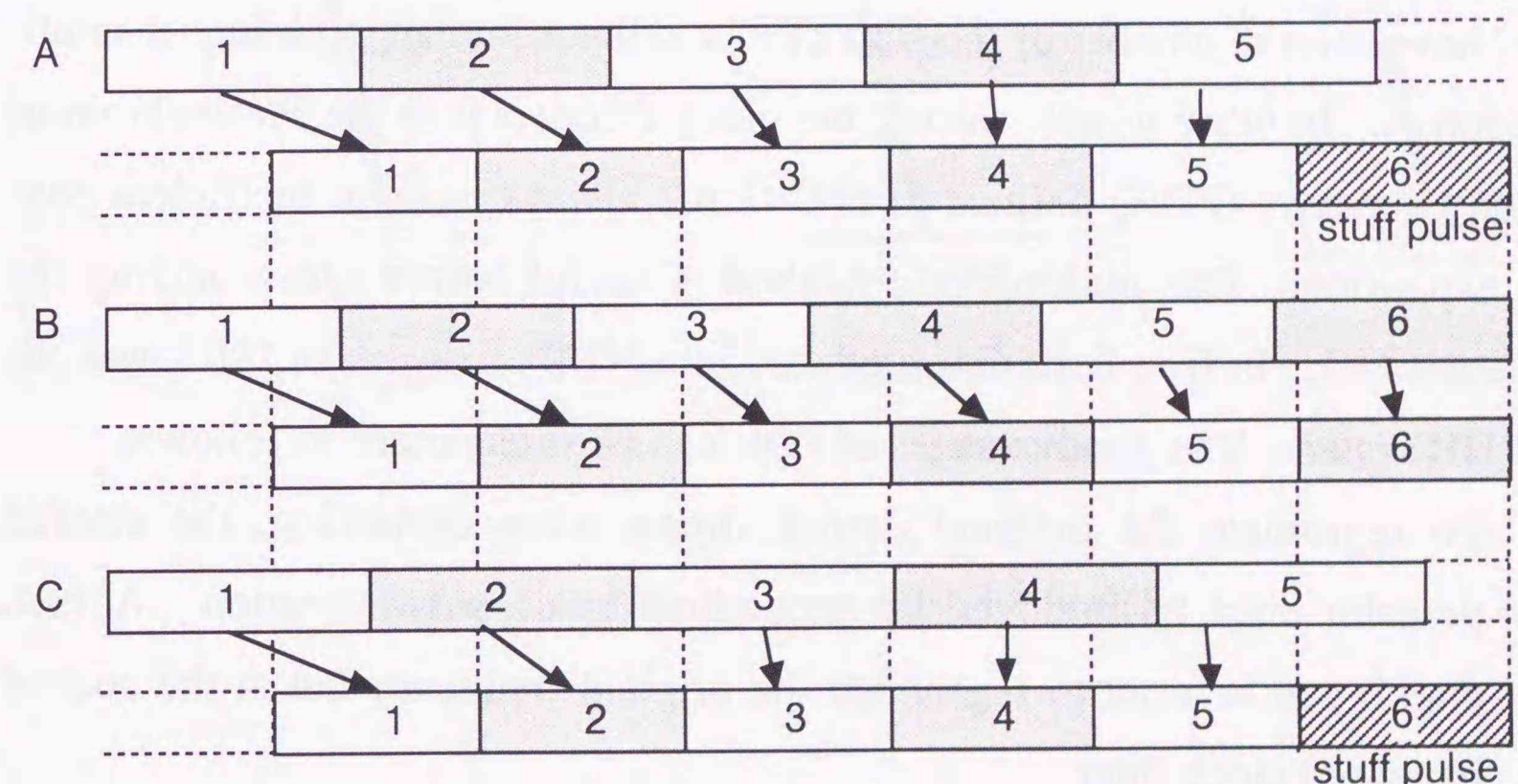


Fig. 1-3 Signal frequency unification onto a faster clock

higher than any of the incoming ones, as shown in Fig. 1-3. The residual problem is also indicated in the figure as two hatched pulses (#6 pulses in top and bottom unified streams), which carry no information. These empty timeslots are designed to carry "stuff pulses". At the receiving port, these stuff pulses have to be removed. To do this, the occurrence of stuff insertion is indicated in the designated timeslot, which is multiplexed within the data pulses. The demultiplexing terminal removes those stuff pulses, only if the stuff indicator is on, and then reproduces the original streams.

Further multiplexing levels can be handled in the same manner as described above. Namely, stuff multiplexing is repeated in the higher stage to multiplex the pulse streams which have already been stuffed. Stuff multiplexing network results in a step by step multiplex configuration to reach the higher bit rate stage. Figures 1-4 and 1-5 depict the hierarchy structure and the system equipment. This scheme is advantageous in preventing the inflation of multiplexer circuit scale, however, the number of the stuff pulses will accumulate with higher multiplex stages. In other words, raising the clock frequency in the Plesiochronous Digital Hierarchy (PDH) defined in ITU-G.702 always leads to inefficient payload expansion. This multiplexing method is useful before synchronizing the whole network. In fact, F-1.6G, introduced into NTT's network in 1987, was the last PDH system. The disadvantages of PDH can be summarized as follows.

- (1) To reproduce the original digital stream after destuffing, the stuffed timeslot must be removed by expanding "real" pulse duration. A PLL circuit can be used to regenerate the original frequency but at the cost of increased clock jitter.
- (2) The extraction of constituent signals in the aggregate bit stream at an intermediate site for dropping and adding tributaries necessitates going through the whole process of demultiplexing down to the appropriate lower level and remultiplexing to the line rate [22]. This makes it quite impractical to access constituent channels in the multiplexed line signals.

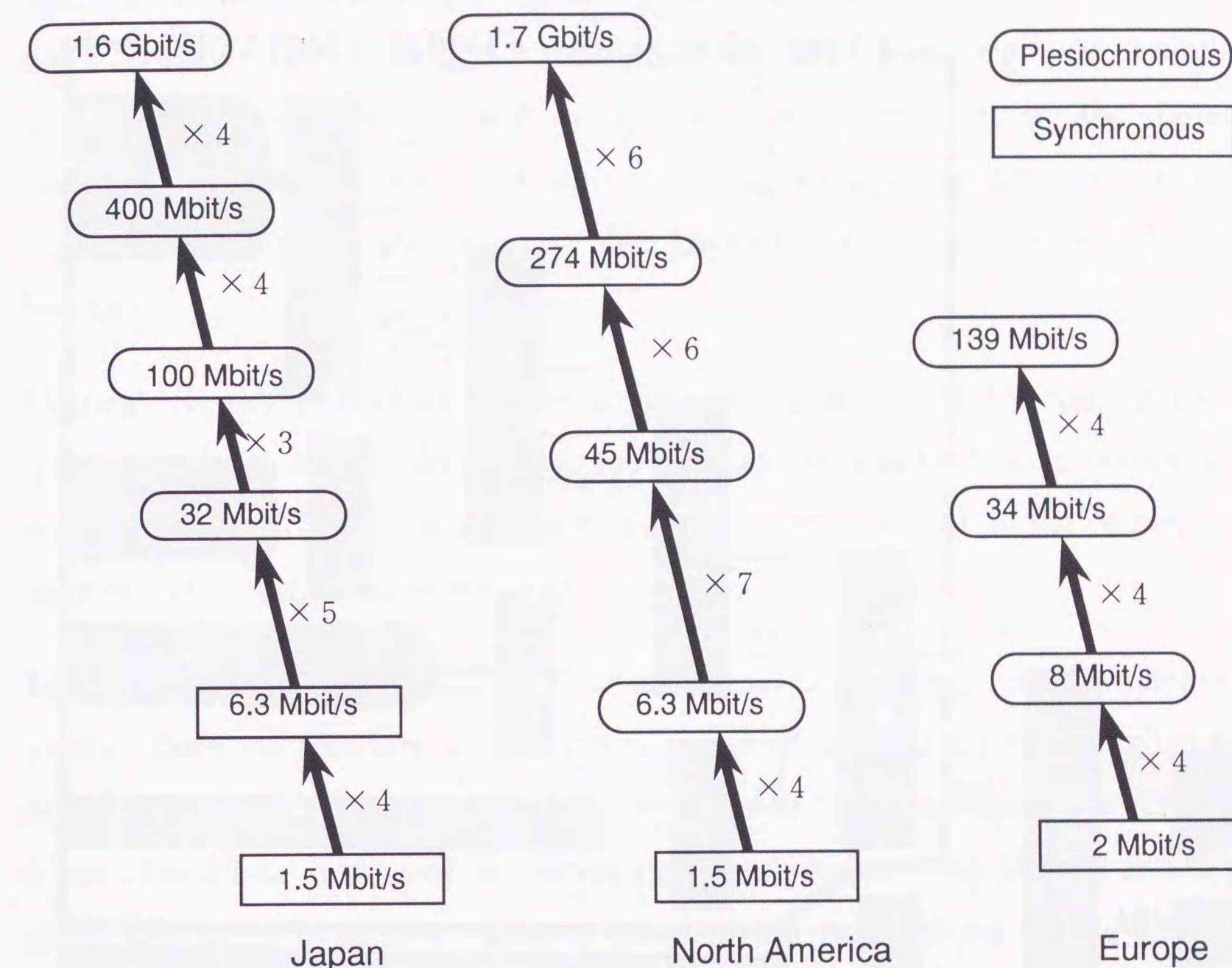


Fig. 1-4 Plesiochronous digital hierarchy

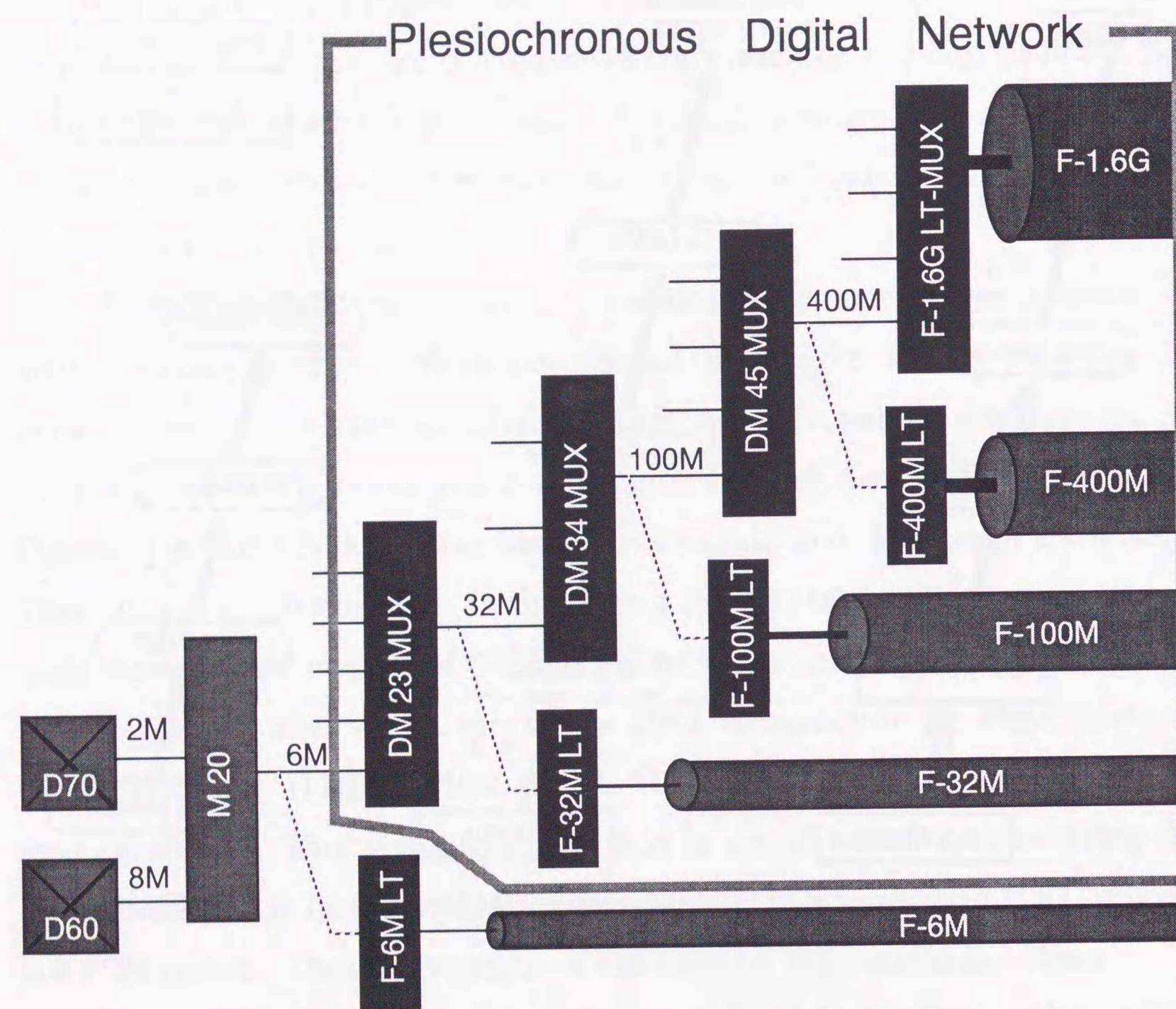


Fig. 1-5 Equipment system configuration for the PDH network

1.4.2. Synchronizing a network [23]

Master-slave Synchronization: The master source generates a clock signal of sufficient quality. In SDH, this signal is distributed via the transmission network, not a separate network. Slave clock regenerators, which are deployed in almost all communication sites, smooth out the phase noise imposed by the signal's distribution, reproduce the clock signal, and supply it to the dependent equipment. To improve reliability, a complete backup master is placed in a different location.

Mutual Synchronization: A clock generator compares all incoming clock frequencies with its own frequency, and feeds the results back to the clock network. Since malfunction in a clock can affect the accuracy of all the others, it is relatively difficult to locate the fault.

Independent Synchronization: No clock signal is exchanged between subnetworks. Only the accuracy of each clock generator prevents a frame "slip" at the boundary between the subnetworks. "Slip" means that the receiving terminal misses the frame timing of the incoming signal, mainly due to data errors or clock frequency mismatch. Recent international connections are based on a technique that achieves 10^{-11} frequency accuracy.

1.4.3. The early stage of Network Synchronization [24]

Network synchronization was developed so as to overcome the disadvantages of PDH. To do this, master-slave synchronization has been introduced, wherein a high-quality master clock is distributed nationwide to lock the slave's frequency. Propagation delay fluctuates partly due to the cable length changes caused by temperature changes, and to parameter changes in the timing circuits. However, the average frequency is identical across the network as long as locking is maintained. Since the phase change of the tributary signals can be absorbed in

buffer memories, signal phase can be kept aligned. More concretely, the first action is to accumulate one-frame data into memory, then read it out to the next multiplex frame to be transmitted. This procedure holds the maximum delay in multiplexing to one-frame. This delay, which cumulates at every synchronous multiplexing stage, was a serious problem.

In 1987, the Japanese network was synchronized up to the second stage (6.3 Mbit/s), which is equivalent to 96 telephone channels. The same capacity stage in North America was not synchronized at that time. In Europe, only the first stage, 2 Mbit/s capacity, has been synchronized, as shown in Fig. 1-4.

1.4.4. Multiplexing in the Synchronous Digital Hierarchy

Synchronous Digital Hierarchy (SDH) multiplexing is an extension of conventional synchronous multiplexing [25], [26]. ITU has succeeded in reaching a world-wide agreement on a digital hierarchy and interfaces. The open equipment specifications allow international manufacturers to challenge new markets with competitive equipment prices. The goal has been to achieve efficient multiplexing not only of the low speed (bit rate) signals already in service but also of the widest possible variety of high speed services in the future. In November 1988, ITU issued an SDH recommendation which made it possible to connect high-speed signals over international borders. The unified bitrate is now 155.52 Mbit/s (STM-1) and integral multiples, times four (STM-4) and sixteen (STM-16), as shown in Fig. 1-6. One of distinctive features in the synchronous digital hierarchy is that a higher bit rate stage can be directly generated from lower STM-1, as shown in Fig. 1-7. In this paper, the author proposes to add a new higher step, STM-64 or 9953.28 Mbit/s. The signal frame adopted has a 125 μ s period (8 kHz repetition), well suited for voice signal processing. A new "pointer" technique is introduced to avoid delay until the multiplex frame starts. Tributaries are conveyed through an imaginary pipe, called the "path", to the

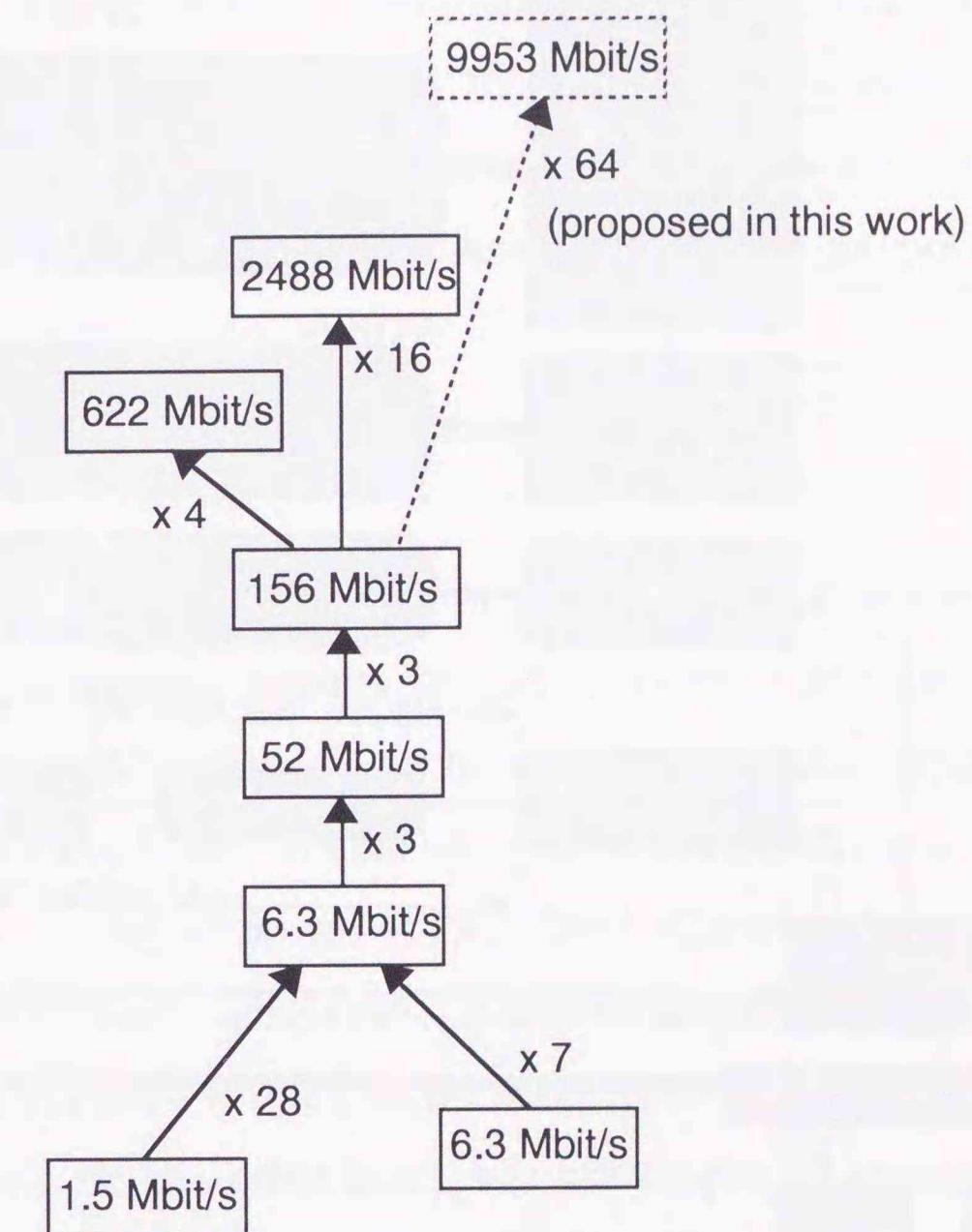


Fig. 1-6 Synchronous Digital Hierarchy

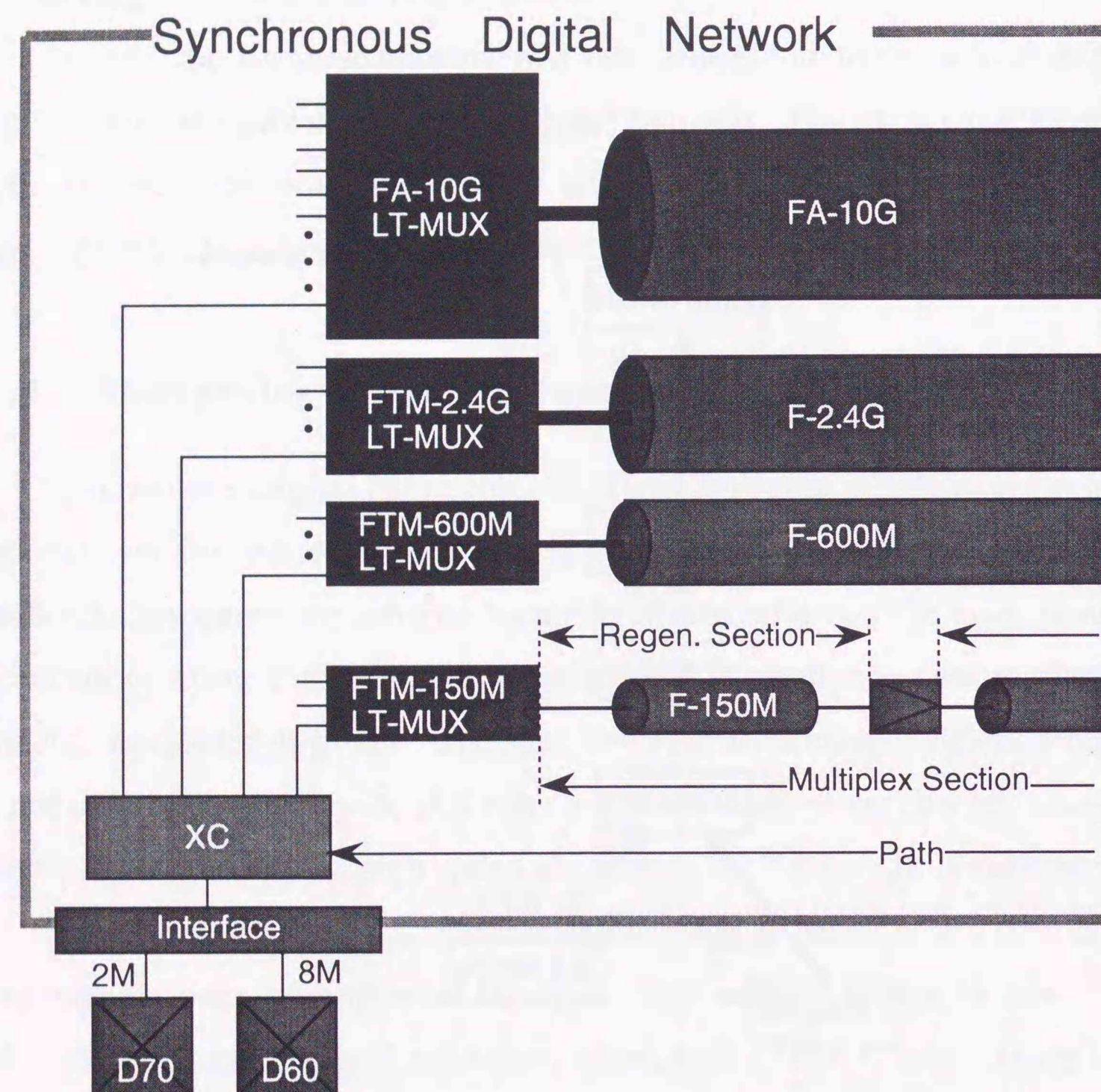


Fig. 1-7 Equipment system configuration for the SDH network

receiving end. In order to allow more customers to share the transmission line, paths traveling in the same direction are multiplexed. Since each tributary is identified by the path overhead (POH) bytes in its beginning, a receiving terminal must detect when the POH arrives. Let us name the set of a tributary and its POH the administrative unit (AU). SDH multiplexing uses a 10-bit "pointer" located within the aggregated frame overhead (Section Overhead [SOH]) to digitally indicate the relative delay from SOH to POH, as shown in Fig. 1-8. The pointer value (= delay) calculation makes it unnecessary to hold the AU until the SOH of the next frame starts. The maximum delay of 125 μ s, possible due to the accumulation, is thus minimized in SDH.

1.4.5. Transmission Coding

In this paper, "transmission line" or just "line" refers not only to the fiber itself but also to the repeater circuits needed to compensate signal attenuation or deformation. Many transmission codes have been invented to suppress the degradation due to imperfect lines. In coaxial cable systems, all input pulses are converted into a code having no power at DC (Direct Current), because DC is occupied by the power delivery subsystem. Balanced ternary codes, taking electrical positive or negative values with equal probability, are widely employed. Optical digital transmission, on the other hand, has not employed ternary coding. The first reason is that the fiber provides a vast bandwidth compared with coaxial cables. Narrow bandwidth occupation, a characteristic of multi-valued codes, is of less importance in fiber systems. Multi-valued codes were considered impractical in fiber systems due to their limited output power and the difficulty of realizing linear modulation. High-speed binary codes demand only simple circuits and are very popular so we confine ourselves to bipolar codes in this paper. In choosing among the many binary codes known, the important points include support of the line watching function, provision of a subchannel for maintenance

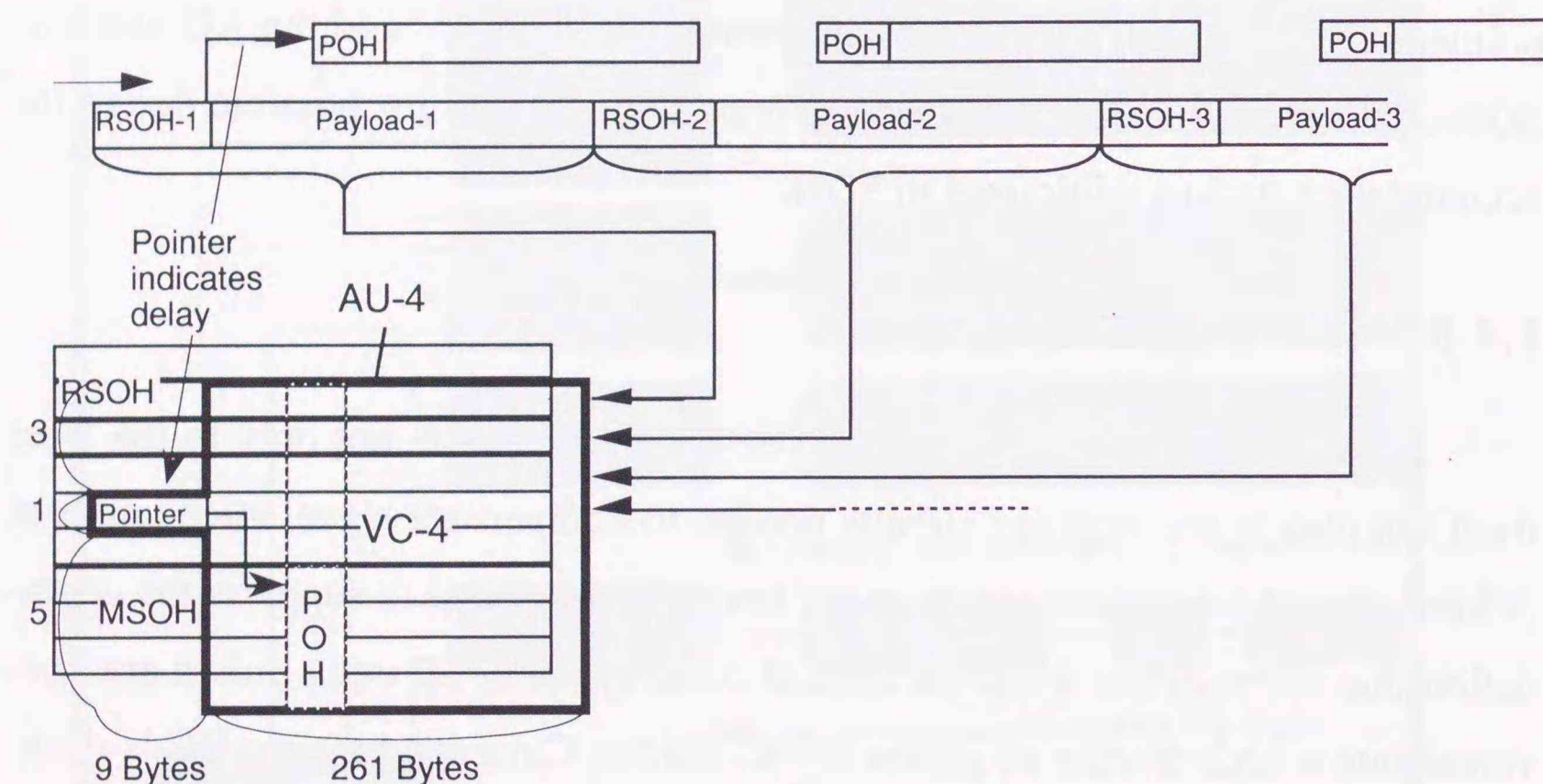


Fig. 1-8 Synchronous Transport Module -1

signals, circuit simplicity, and low-power consumption.

As in many other applications, analog circuits play an important role in equalizing and amplifying the received pulses so as to meet the waveform conditions need for the following digital circuit. The parity of 1's and 0's in the data should be balanced to avoid baseline wander because in many cases there is AC coupling between the photodetector and the subsequent analog amplifier. Such wander in the average power should be also suppressed when the gain-controlled amplifier or output optical level is controlled through the average power of the pulse stream.

A sufficient density of the transitions between 1 and 0 is needed to prevent dying out the signal from the clock extraction circuit, which is composed of a narrow-band filter (timing tank) centered at the clock frequency followed by a limiting amplifier. One measure to prevent clock failure is to limit the number of consecutive identical symbols that does not provide the clock spectrum. While the output level from the tank is proportional to the transition density, the limiting amplifier absorbs the level variation and keeps the output clock level constant. Imperfection in the amplifier causes clock jitter due to the transition density fluctuation. From a hardware availability viewpoint for high-speed transmission, codes are more favorable if they require less clock increase and can be realized as simple circuits.

The above functions are accomplished through the use of scramblers, bit insertion coders, or block coders.

Scrambling: Scrambling is used to randomize input patterns by adding it modulo 2 (exclusive-or). This reduces the possibility of generating a long string of identical symbols. A self-synchronous scrambler-descrambler with five stages, is shown in Fig. 1-9. This process does not need resetting. Simply feeding in the data will result in the output of a scrambled string after the registers fill up.

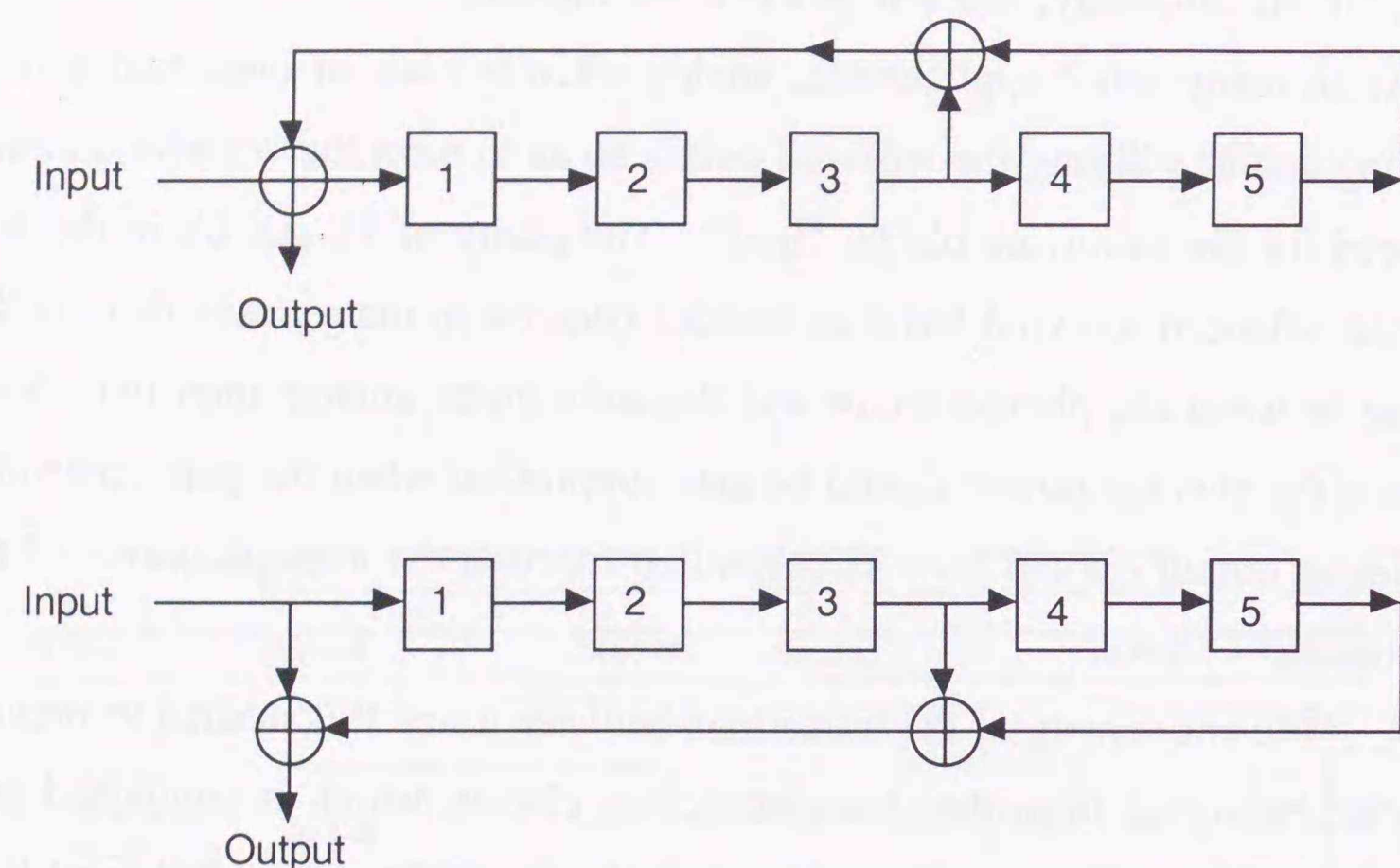


Fig. 1-9 Self-synchronous scrambler (top) and descrambler (bottom).

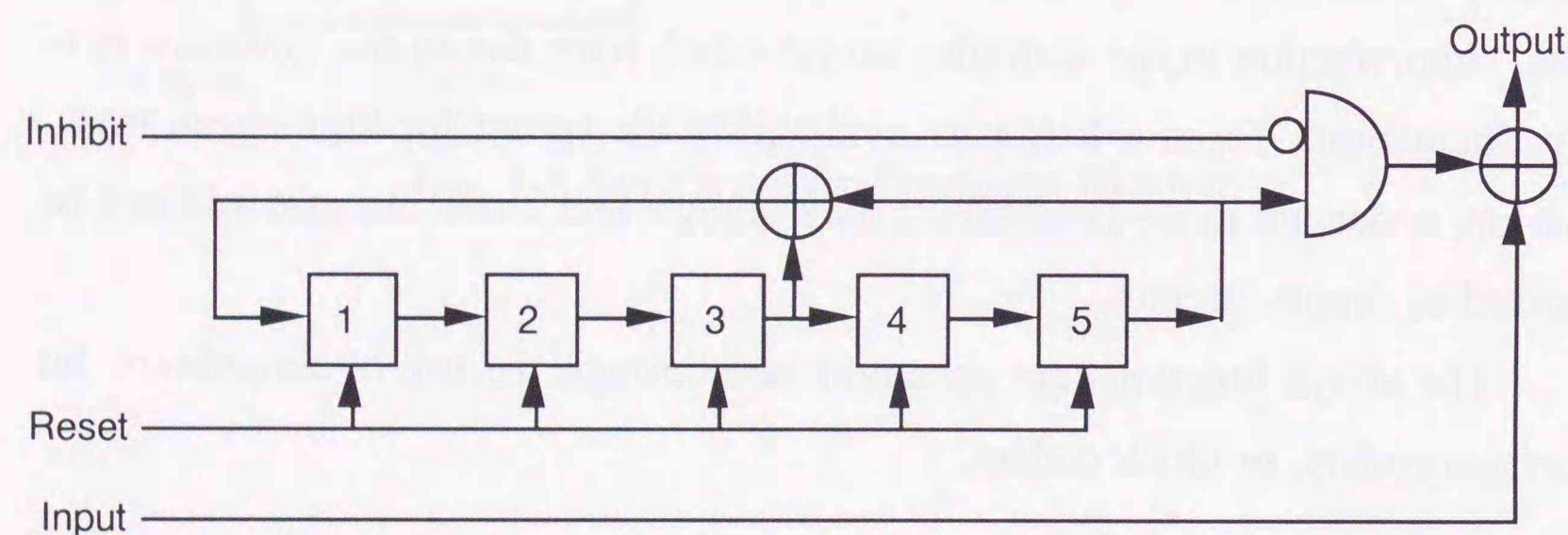


Fig. 1-10 Reset-type scrambler/descrambler

Only a simple circuit is needed, however the characteristic of error-expansion has to be considered. For example, imagine that one bit error arrives at the descrambler shown in the bottom of Fig. 1-9, the following third and fifth bits are also corrupted. One-bit error is tripled in this type of descrambler. Another type is the one with reset, shown in Fig. 1-10, where the circuits for the scrambler and the descrambler are identical. The descrambler must detect the reset timing from the incoming data stream. Frame timing detected by the framing circuit can be easily transformed into the reset pulse, provided that the framing bytes are kept unscrambled. The F-1.6G system uses a seven-stage reset type scrambler in conjunction with 10B1C bit-insertion coding. The SDH standard adopts ten-stage type scrambling.

Bit Insertion Coding: If we read out an n bit data block from memory with a clock $(m+1)/m$ times higher than the writing clock, one open time slot appears at the end of the n bit block. mB1C coding inserts into that slot a bit that is complementary to the last bit of the block. This gives at least one transition within the encoded $(m+1)$ bit [27]. The resulting bit rate increases range from 10%, in the case of 10B1C code, to 100% for dipulse codes. Bit insertion coding can also be achieved with a relatively simple logic gate scheme, so it is attractive in high-speed transmission. Many transmission systems have adopted the mB1C code. Both F-400M and F-1.6 systems employ the 10B1C code.

Block Coding. Block coding requires more complex logic than bit insertion, however, it constrains the code words more strongly in terms of the maximum run length and the maximum digital sum variation. Generally speaking, in nBmB coding, a single error may force the encoded word to be decoded to a completely different n bit block, thus the error expansion is "n". It should be noted that Widmer and Franaszek have reported in an 8B10B paper that expansion can be reduced if the coder is partitioned into 5B6B and 3B4B subordinate ones [28]. Another disadvantage in block coding is that the complexity of the logic circuit

tends to increase rapidly as the coding efficiency or "n" is increased. This prevents its straightforward implementation in high speed systems, where code efficiency is an important parameter in selecting a code.

Line watching is important for detecting symbol errors. As an example, we consider 10B1C coding, where every 10 bit information block should be followed by an additional bit that is complementary to the preceding bit. If we watch the coding rule, some (1/11) errors can be monitored. In the F-1.6G system, rule monitoring is used for fault location, while a parity check is used to monitor the whole multiplex line. In SDH, on the other hand, N bit parallel parity check, or Bit Interleaved Parity-N, is performed against all bits at the sending and receiving equipment. Calculation result at the sending side is written into the SOH byte. It is compared with the one obtained at the receiving side to detect errors. BIP-24 is used for monitoring the entire multiplex section, while BIP-8 is applied at each regenerator section for fault location.

1.5 Multiplex transmission system construction

General configuration of a digital telephone network is shown in Fig. 1-11. The exchange equipment (E) switches calls between dependent telephones and is connected to a regional cross-connect (R-XC) for calls between exchanges. It hands over the signals to the cross-connect (XC) for long-distance communication as shown in the top of Fig. 1-11. The cross-connect system selects available transmission lines and disconnects those that are in trouble. Trunk transmission systems have multiplexing terminals at both ends of the multiplex section, as shown in the bottom of Fig. 1-11. A protection line is prepared against the case that it is necessary to disconnect the working line. The working and protection systems are generally mounted on fibers packed into the same cable. Although the protection switch can quickly recover from fiber cut or repeater circuit failure, it fails if the whole cable fails. The cross-connect system can deal with

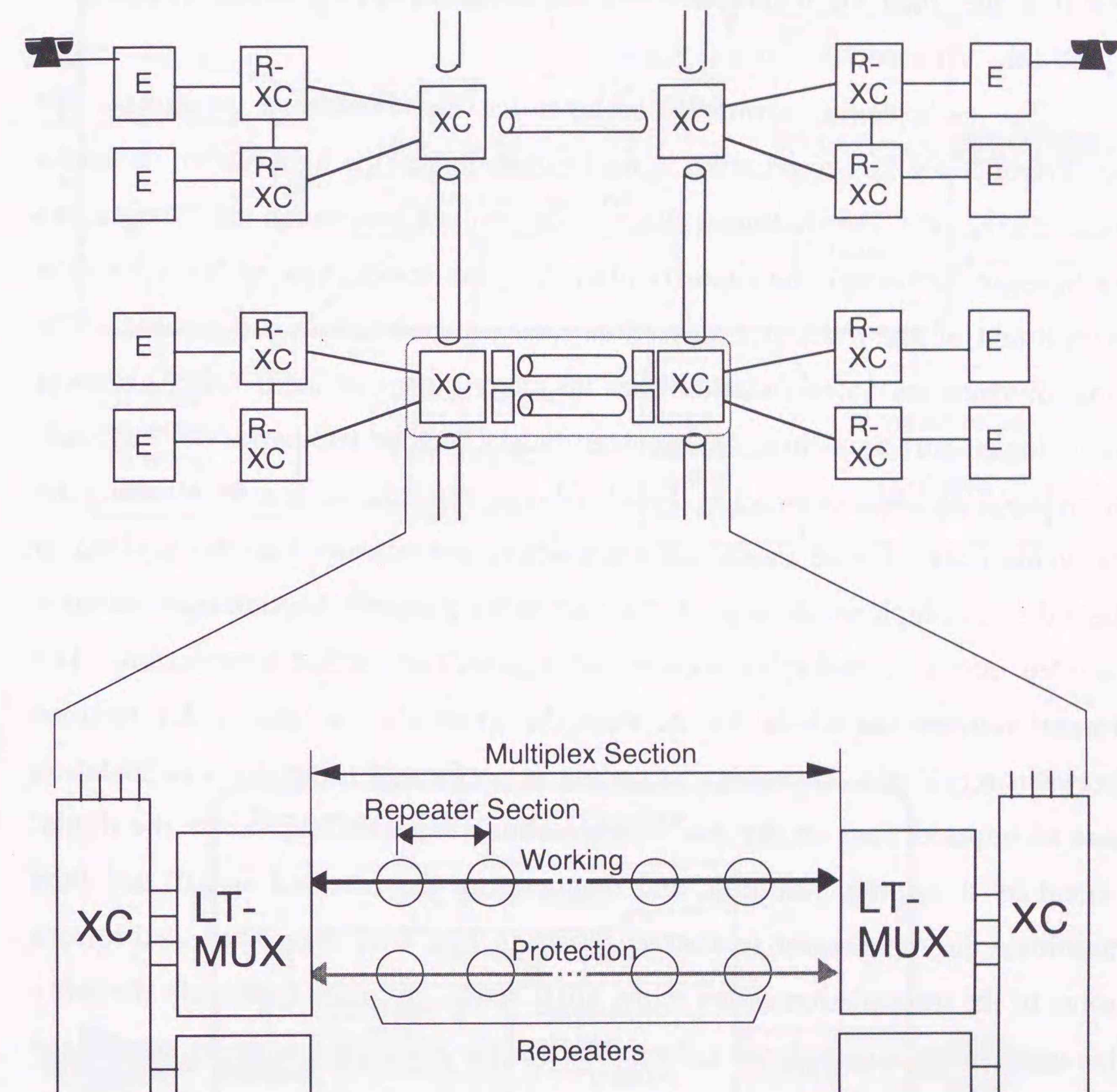


Fig. 1-11 General configuration of a digital telephone network. The exchange equipment (E) switches calls between dependent telephones and is connected to a regional cross-connect (R-XC) for calls between exchanges. It hands over the signals to the cross-connect (XC) for long-distance communication (top). The cross-connect system selects available transmission lines or disconnects those that are in trouble. Trunk transmission systems have multiplexing terminals at both ends of the multiplex section (bottom).

such route disconnection, however, restoration takes longer than is true with line protection switching. Repeater(s) have to be deployed if the length of the medium is longer than the regenerator spacing designed for the multiplexed signal speed (bits per second).

The multiplexing terminal combines the input channels, terminates the section, encodes the transmission signal, and modulates the light carrier, as shown in Fig. 1-12 (a). The multiplex function determines how much information can be transmitted through the capacity offered by the lower physical layer function constituted by the transmission medium and transmitter/receiver circuits. The time division multiplex function lines up many pulses of input lower channels into single short timeslots. Multiplexed pulses include tributaries or "Payload" from the cross-connect switches that choose the transmission line from among the available ones. Frame pulses are multiplexed periodically into the payload to help the demultiplexer distinguish the start of the payload. Section termination is divided into two: multiplex section and regenerator section termination. The former watches the whole line between the terminals, the latter is for sections between repeaters. Transmission coding is performed to suppress degradation due to imperfections in the line. Regenerating repeaters reproduce the digital signal by reshaping, retiming, and regenerating the received signal, and then terminate the regenerator section as shown in Fig. 1-12 (b). They also refresh some of the transmission codes as the mB1C code. In order to identify the problematic section, supervisory information for each section has to be sent to the receiving end. The demultiplexing terminal performs the reverse functions to reproduce the lower channels.

Chapter II describes STM-64, a recently developed 10 Gbit/s multiplexer architecture that is a natural expansion of the SDH recommendations made by ITU. Chapter III details regenerator section termination LSIs intended for the STM-64 system. The terminal function also adds supervisory pulses to monitor the signal quality or to detect circuit failures, and inserts maintenance pulses to

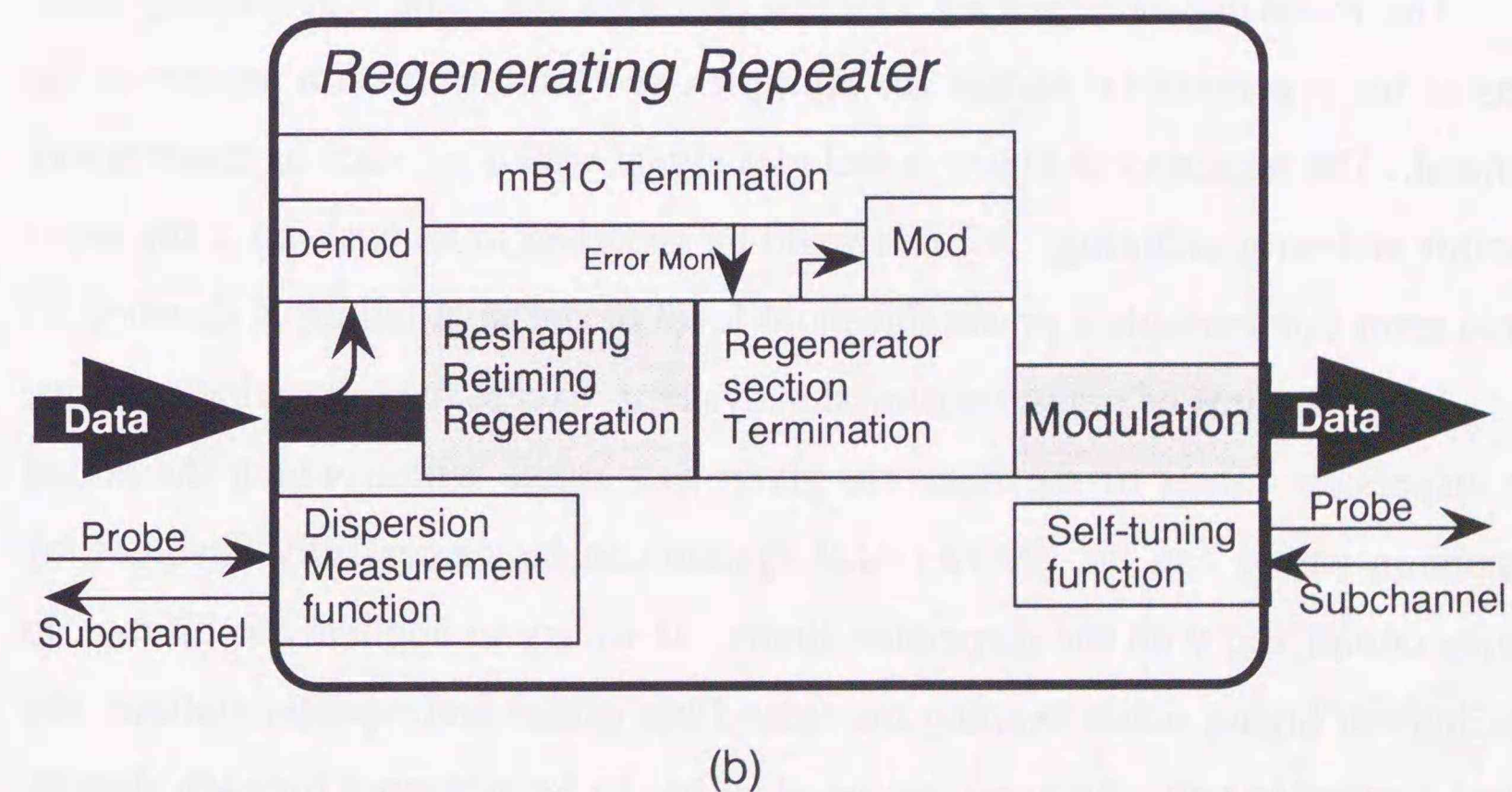
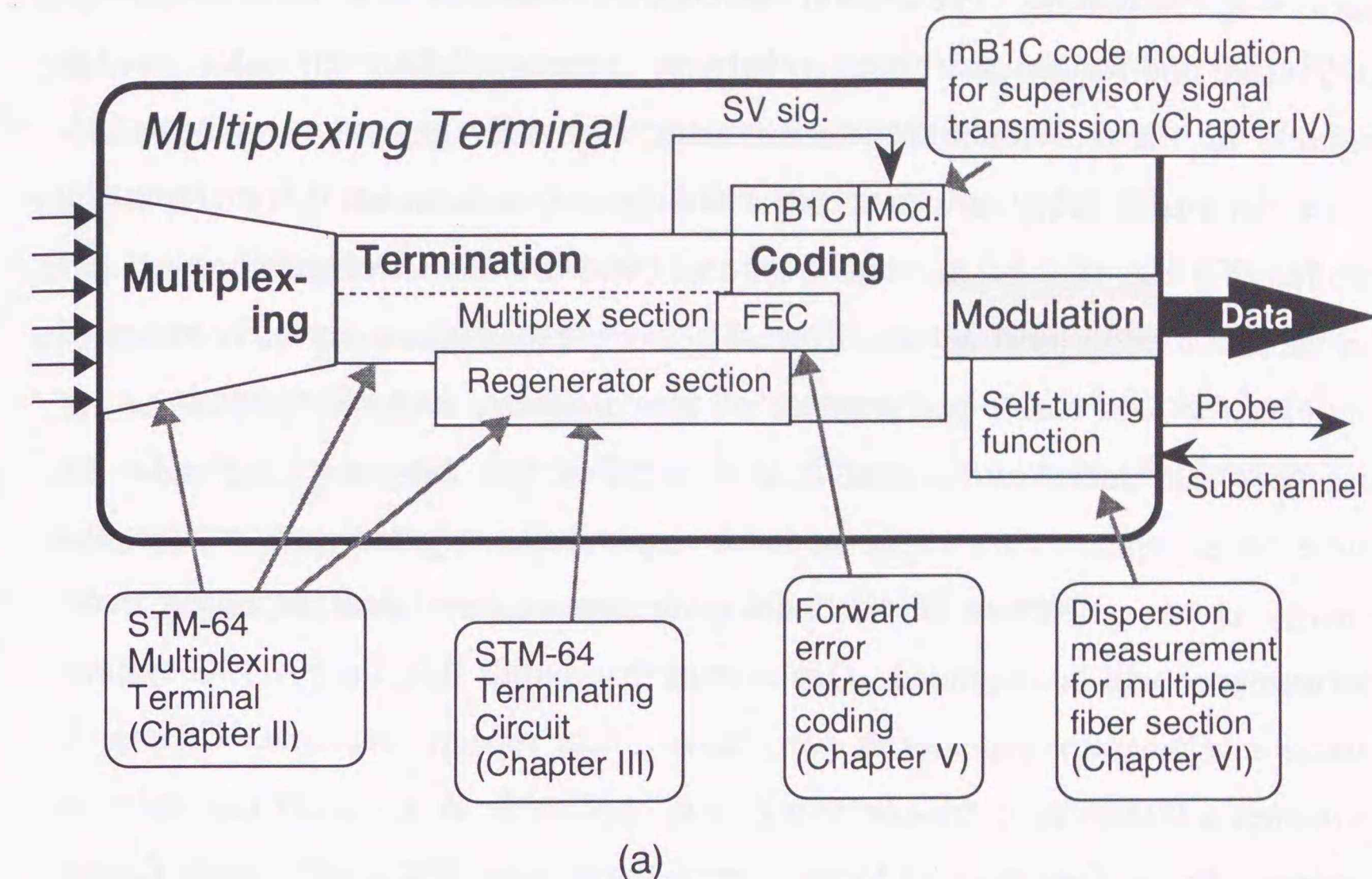


Fig. 1-12 Functions of multiplexing terminal (top) and regenerating repeater (bottom).

control line switching. The section surveillance function can be categorized into two types: one for the entire section between the multiplexing terminals, and the other is for the individual sections terminated by the regenerators. Chapter IV describes a trial subsystem in which a maintenance bit stream is transferred by modulating the "C bit" amplitude of the 10B1C code. Designers select the transmission code most suitable for each given transmission line. By virtue of the excellent and stable performance of optical fibers, no error-correcting code has been introduced into commercial systems so far. However, in the future, more linear repeaters are expected to be employed to suppress construction and energy costs. Because of the analog nature of the linear repeater, error performance will be degraded. One system that avoids this, the FA-10G system, places a regenerator repeater at every three linear repeater sections. Chapter V proposes a Hamming error-correcting code applicable to any SDH transmission system. The introduction of forward error correction (FEC) will enable further cost reduction of such transmission systems, while keeping frame compatibility with uncoded systems.

The receiving multiplexing terminal performs the same regenerating functions as the regenerators, except for multiplex-section termination unique to the terminal. The termination function includes signal watching, such as input power monitor and error counting. A line should be switched to its back-up if the monitored error rate exceeds a preset threshold level or decisive failure is detected.

In the designs of many transmission systems, except for submarine systems, the dispersion values of the fiber are given as a range within which the actual dispersion values can lie. A 10 Gbit/s system can be successfully designed by simply complying with the dispersion limits. If we try to upgrade the bit rate, to 40 Gbit/s or higher while keeping the same fiber cables and repeater stations, the actual dispersion value between regenerators has to be measured for each system. In order to realize fully automated dispersion measurement, chapter VI proposes a method applicable to multiple-fiber sections, even if narrow-band optical

amplifier repeaters are used between sections. The method requires just a single phase-modulated (PM) light source and relative amplitude detection is sufficient to measure the dispersion. If a transmitter scans a wave-length window with the tunable PM light, the receiver terminal can determine the optimum wavelength from the amplitude modulation data recorded. Upon being informed through a maintenance subchannel, the transmitter locks on to the optimum wavelength. This self-tuning experimental system has already been successfully tried and a report has been accepted for OFC'97 [29]. Chapter VI gives the dispersion measurement principle of this innovative system.

Chapter VII contains the concluding remarks of this paper.

Chapter II

Line Terminating Multiplexers for SDH Optical Networks

2.1. Introduction

The Integrated Services Digital Network (ISDN) is being developed to provide a wider variety of services, such as computer communication, and video transmission than is currently available. High-speed synchronous networks are fundamental if such broad-band services are to be offered. Network synchronization is essential for economical network construction and operation. Multiplexing costs can be reduced because network synchronization permits the direct multiplexing, instead of step-by-step multiplexing, of not only lower channels, but also the Operations, Administration, and Maintenance (OA&M) channel. NTT has developed the STM-4 (622 Mbit/s) and STM-16 (2.4 Gbit/s) systems in accordance with the Synchronous Digital Hierarchy (SDH). "STM" stands for the synchronous transport module in the synchronous digital hierarchy

(SDH) defined by International Telecommunication Union (ITU). The systems, however, fail to fully utilize the broad-band capacity of single-mode fiber. The need to achieve further increases in communications capability is propelling the current research trends. Direct optical amplification has enabled multigigabit-per-second signals to be transmitted over several hundreds of kilometers without regeneration [30]. Although wide-band electronic amplifiers capable of amplifying picosecond pulses are extremely difficult to make, the optical amplifiers based on single-mode silica fiber doped with erbium ions have been confirmed suitable for 20 Gbit/s signals [31]. Erbium doped fiber amplifiers (EDFAs) can be used as boosters, linear repeaters, as well as preamplifiers, and have made multigigabit transmission systems more feasible. There is, however, another technology on which multi-gigabit-per-second transmission systems may be developed. Wavelength Division Multiplexing (WDM) offers the unrestricted multiplexing of channels that differ in terms of their signal format and signal rate. Optical Frequency Division Multiplex (OFDM) may imply denser channel spacing, but this paper treats OFDM as just one reference to WDM. WDM must be included in any comprehensive review of promising optical transmission equipment.

This paper analyses the impact of various SDH requirements on the design of line-terminating multiplexers (LTMs). We compare WDM and two types of Time Division Multiplexing (TDM), namely the interleaving unit is either a bit or byte. The construction of an LTM for a multigigabit-per-second system is then examined. A simple byte-interleaving circuit configuration is introduced to show that multiplexing multiples of eight channels is most advantageous in terms of circuit simplicity [32]. A prototype STM-64 LTM is developed with the emphasis on lower channel termination, repeater-section termination, and byte interleaving at up to 10 Gbit/s.

2.2. Comparing Multiplexing Methods; WDM or TDM?

Three multiplexing methods are possible to construct higher-capacity transmission lines on SDH networks. They are Wavelength Division Multiplexing (WDM) and two types of Time Division Multiplexing (TDM): bit interleaving and byte interleaving. Bit interleaving is the only choice for plesiochronous networks, because byte interleaving is virtually impossible. Only in synchronized networks is byte interleaving possible, and thus it (byte interleaving) is employed in SDH Network Node Interfaces (NNIs) [33].

2.2.1. Disadvantages of WDM

In the WDM system, each wavelength channel should be allocated a unique wavelength and the wavelength spacing must be regular, as shown in Fig. 2-1(a). Unfortunately, fiber dispersion and loss usually depend on the wavelength. The total bandwidth occupied by a WDM system must be wider than that occupied by its TDM counterpart carrying the same capacity, because WDM needs guard bands between channels to minimize crosstalk. In WDM systems, it is favorable from the view point of fiber nonlinearity to use a high-dispersion wavelength region to reduce the signal degradation caused by the Four-Wave Mixing (FWM) process [34]. This means individual channel capacity is limited by dispersion unless its influence can be equalized. It is obvious that unequal separation lessens the influence of FWM [35], however suitable optical multi/demultiplexers would be extremely difficult to fabricate. In TDM systems, the spectrum can be allocated to the low- or even zero-dispersion wavelength if it is preferable, as shown in Fig. 2-1 (b).

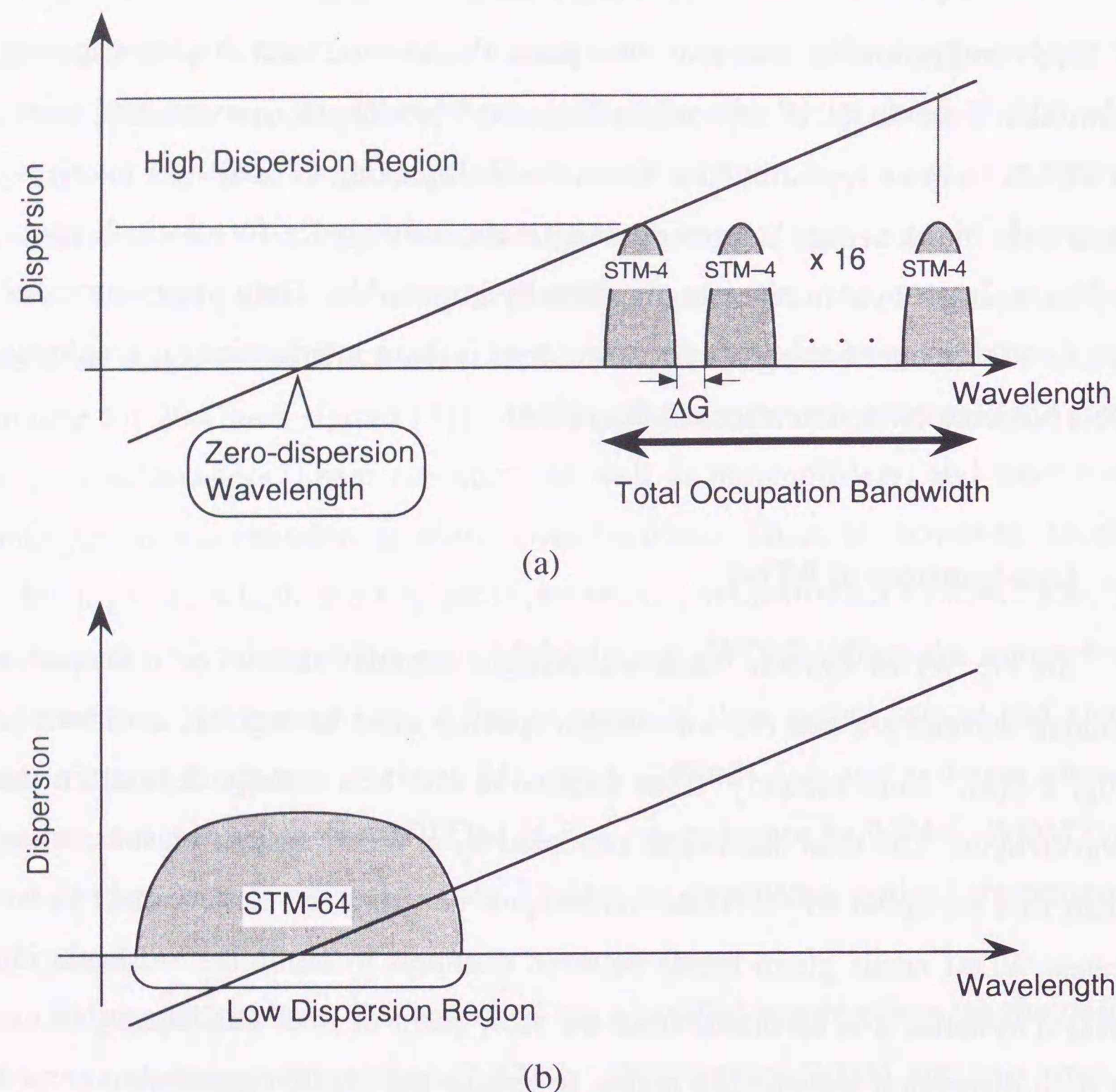


Fig. 2-1 Spectral allocations for the (a) WDM and (b) TDM systems.

Another possible problem is the different group velocity of each wavelength channel. In other words, WDM channels may not be synchronized to each other. In principle, TDM systems allow the maximum input capacity to equal the line rate. This is not automatically guaranteed in the WDM system. Assume a wavelength window near the zero-dispersion window of a transmission fiber, and the following parabolic dispersion curve;

$$y(\lambda) = \left(\frac{a}{2}\right)(\lambda - \lambda_0)^2 \quad (2-1),$$

where λ_0 is the zero-dispersion wavelength that maximizes the group velocity. The parameter a is the slope of the dispersion curve, and is at most 0.1 ps/(nm² km) for commercial fibers. We also assume two wavelength channels;

$$\lambda_i = \lambda_c \pm \frac{\Delta}{2}, \quad (2-2)$$

where $i = 1, 2, \dots$. The relative group delay per fiber length for the two wavelength channels should be;

$$y(\lambda_1) - y(\lambda_2) = a\Delta(\lambda_c - \lambda_0). \quad (2-3)$$

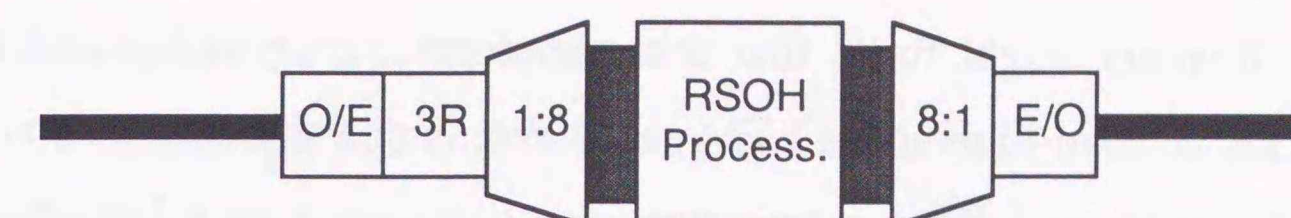
We assume the severe situation in which the central wavelength λ_0 is located 15 nm from the zero-dispersion wavelength. To realize a capacity of 10 Gbit/s by multiplexing four STM-16 channels with a separation of 1 nm, the maximum wavelength difference becomes 3 nm. By substituting these values into Eq. (3), and taking a repeater spacing of 80 km, the relative total delay is 360 ps. This occupies 90% of the STM-16 unit interval (400 ps). This result implies that the WDM channels considered here are unsynchronized, unless an automatic delay detection and compensation function is employed. At present, this is virtually impossible in a practical transmission system. Precise control of both WDM filter and laser source wavelength and their sharpness can

narrow the WDM spacing and so overcome this difficulty; however, the technologies are still being studied [36]. Therefore, it is concluded that WDM is too immature to realize ultra-large capacity systems even though it does not require any high-speed electronics to multiplex and demultiplex the optical signal.

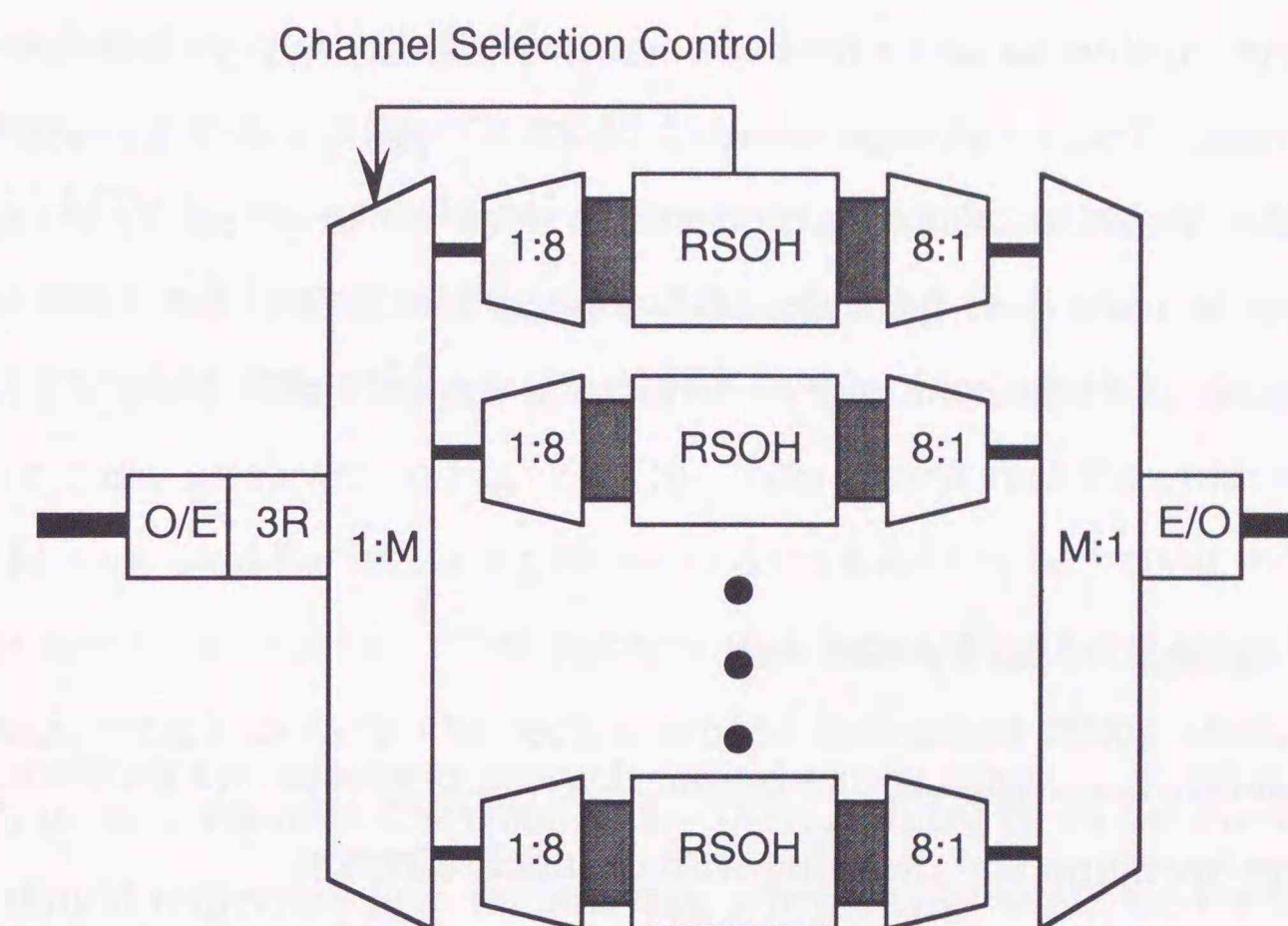
2.2.2. Bit Interleaving or Byte Interleaving?

Bit interleaving might seem to be advantageous because its circuits appear to be much simpler than those demanded by byte interleaving. The former simply multiplexes input bits from one channel to another. However, after furnishing the equipment needed to realize the essential SDH functions, one has to accept that bit interleaving is, in fact, too complex. For example, bit interleaving requires the incoming bits to be synchronized, not bytes. We can not expect, however, that bit synchronization is perfect in a synchronous network. "Synchronous" means that the average clock rate is equal for all equipment. All equipment has to be designed to accept some small, but significant, phase difference in the signals as received. Consequently, a frequency justification process is required to adjust their incoming clocks to the internal reference. One solution is the pointer operation used by the SDH-NNI.

Systems connected in the near future to the existing SDH network must also support overhead termination. The error monitoring capability realized with Bit Interleaved Parity codes (BIP-8, -24) enables automatic and quick fault location. If a transmission line does not support such a scrupulous monitoring function, an alarm indication generated within the line would propagate until it is terminated. A network management function would be needed to determine the location of the malfunction or disconnection. Self-healing capability would be difficult to achieve in such systems. For this rea



(a)



(b)

Fig. 2-2 Comparing regenerator configurations:
(a) STM-L (L = MN) byte interleaving regenerator.
(b) M channels STM-N bit interleaving regenerator.

son, SDH termination is also necessary in future LTMs, either with bit or byte interleaving, if they are connected to the SDH network.

The difference between the two forms of interleaving is that byte interleaving needs signal transformation to regenerate byte-based sequences after multiplexing. Readers might think that this transformation is redundant when implementing an ultra-high speed LTM. However, since signal functions such as frame detection and error monitoring, are performed on a byte-by-byte basis in SDH equipment, a simple one-to-eight demultiplexing format directly supports the functions with the byte-parallel format. Bit interleaving on the other hand, requires frame alignment and other repeater section overhead (RSOH) processes to be performed after demultiplexing which reproduces the input bit rate. The most significant difference appears in the regenerator circuit scale. A bit interleaving regenerator for M channels of STM- N is M times larger in scale than the byte interleaving equivalent if the transmission signal consists of a byte-multiplexed STM- L ($L = MN$) signal as shown in Fig. 2-2.

2.2.3. Disadvantages of Bit Interleaving

This subsection explains that bit interleaving produces two problems: one with system operation and the other with channel selection.

Consider a TDM system that uses bit interleaving. Line monitoring has to be done at the lower input bit rate after demultiplexing. This leads to the situation wherein each equipment outputs independently a number of error indications for each channel. An interfacing process is needed to realize a unique and consistent alarm from the monitoring signals. Furthermore, if the frame phases of the demultiplex channels are not in phase, the processor has to wait until the monitoring signals of all channels are ready for processing. These impositions mean that bit interleaving systems require more complex

processes for error handling and system control than supported by the standardized SDH equipment. This problem is also true for the WDM system. Therefore, multiplexing schemes that do not use frame phase synchronism, such as bit interleaving or WDM, require very sophisticated, *i.e.*, costly, monitoring and control schemes.

Bit demultiplexed channels may not be always transferred to the right ports without some control technique, such as adding another overhead frame onto the multiplexed signal. However, the additional bytes increase the clock rate required. This increase may not be supported by ultra-high speed electronics operating at or near their limits. There are, however, other ways of distinguishing one channel from another after demultiplexing, such as uniquely numbering the "C1 byte" over all multiplexed channels, or inserting different frame bytes for each channel. The channel selector at the receiving side will rotate the demultiplexed channels until the right one emerges from the right port. The channel selector should be controlled by the frame detection circuit, as shown in Fig. 2-2 (b). This means that the control signal will not be sent until the incoming frame is determined to be wrong, which takes at least one frame period. This ensures that frame alignment takes at least two frames, which exceeds the recommended maximum frame alignment period (250 μ s or 2 frames) described in G. 783. Judging from the above problems, we should utilize the byte interleaving scheme only while the repeater's analog circuit is functioning normally. The WDM technique will be important when the total bit rate exceeds the electrical circuit performance. Therefore, the following discussion will focus on implementing byte interleaving in ultra-high speed transmission equipment. The method proposed herein was designed for byte interleaving. This method offers significant benefits for constructing an LTM when multiplexing integrals of eight channels, that is, STM-8, -16, -24, and so forth. A prototype LTM designed for STM-64 multiplexing is also introduced.

2.3. "STM-64" FRAME ISSUES

2.3.1. SOH Utilization Efficiency

In the STM-N format, one negative aspect in extending the scale of multiplexing is that the utilization efficiency of the section overhead (SOH) bytes decreases. Efficiency is defined here as the ratio of the number of bytes assigned for certain purposes to the total number of SOH bytes. The STM-N frame can be thought of as a stack of basic STM-1 frames, where the SOH occupation rate is fixed for all STM-N frames at 1/30. However, only the first layer of the SOH is used most efficiently. The utilization ratio is 53% for the STM-1 frame, whereas it is 36, 32, and 31% for STM-4, -16, and -64 frames, respectively. Naturally, the number of undefined bytes in the STM-N frame tends to increase with the multiplexing number. The percentage of undefined bytes to total bytes increases as N increases. The percentages are 1.57, 2.12, 2.26, and 2.29%, for STM-1, -4, -16, and -64, as shown in Fig. 2-2. A comparison is possible with plesiochronous systems, such as F-400M, and F-1.6G, that adopt the "10B1C" code. In the 10B1C code, a bit that is complementary to the last bit is added to every block consisting of 10 bit. Therefore, the additional bit occupancy is 1/11 or 9%, which is relatively large comparing with the additional bytes created in the STM frames mentioned above. Furthermore, there is a possibility of using these SOH bytes for network management purposes [37]. The sophisticated management schemes that will be required in the future will need the network to carry much more SOH information. Undefined bytes would be a great help in constructing such a network.

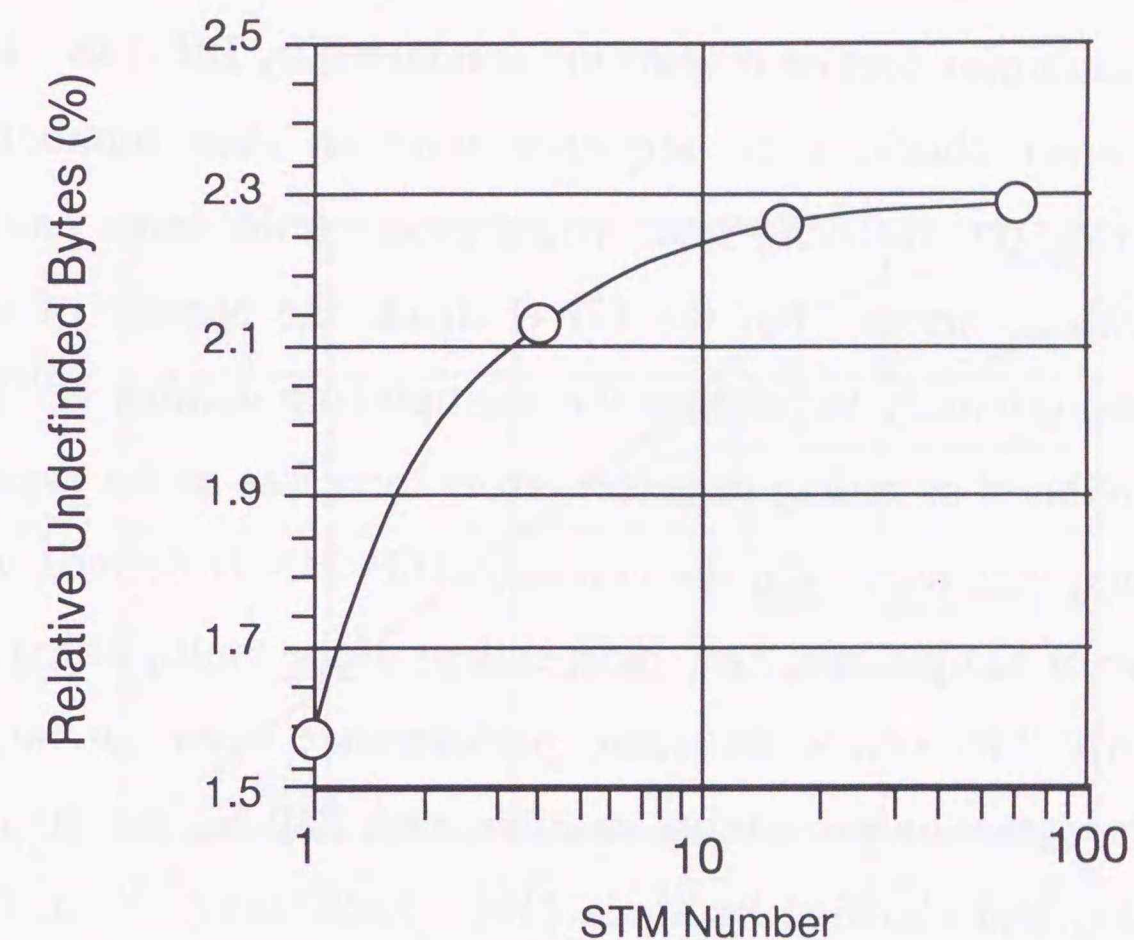


Fig. 2-3 Percentage of undefined section overhead bytes to the total frame bytes.

2.3.2 Bit Interleaved Parity

The ideal bit-error monitor has to output some signal that indicates the number of errors detected. Two error checks are performed in the STM-N frame: one is the regenerator section monitoring performed by BIP-8 and the other is the multiplex section monitoring conducted by BIP-24N. However in any parity error check, a bit sequence with an even number of errors detected during the checking scan would produce the same parity as a bit sequence with no errors. For the BIP-8 check, the number of sample bits becomes proportionally larger than the multiplexing number N . Because of this, the likelihood of failing to detect errors increases as the capacity of the STM-N frame increases. On the contrary, BIP-24N is defined with a constant number of sample bits, *i.e.*, independent of the multiplexing number N ; therefore, BIP-24N offers the same performance even for large N as in STM-1. The detection probability possible with BIP- N_B was first elucidated by Jones [38] and clarified by Ueda [39]. Each parity bit in the BIP- N_B scheme detects a random error on the line with a probability of;

$$M_{BER} = \left[1 - \frac{1 + (1 - 2\varepsilon)^{N_R}}{2} \right] / N_R \quad (2-4)$$

where ε is the actual line error rate, and N_R is the sample bit number assigned to each BIP bit. The error detecting ability for three cases of BIP-8, *i.e.*, STM-4, -16, and -64, are compared in Fig. 2-4. Obviously, BIP-8 for STM-64 has the lowest accuracy. It is unable to accurately measure error rates in excess of 3×10^{-6} . For an actual error rate of 10^{-6} , BIP-8 for STM-64 can detect 85.9% of the errors, whereas STM-16 can detect 96.2% of them. One way of overcoming this difficulty is reducing the number of samples that are checked. For example, replacing BIP-8 with BIP-24 can increase the accuracy to 95.0% for STM-64. The decision whether we

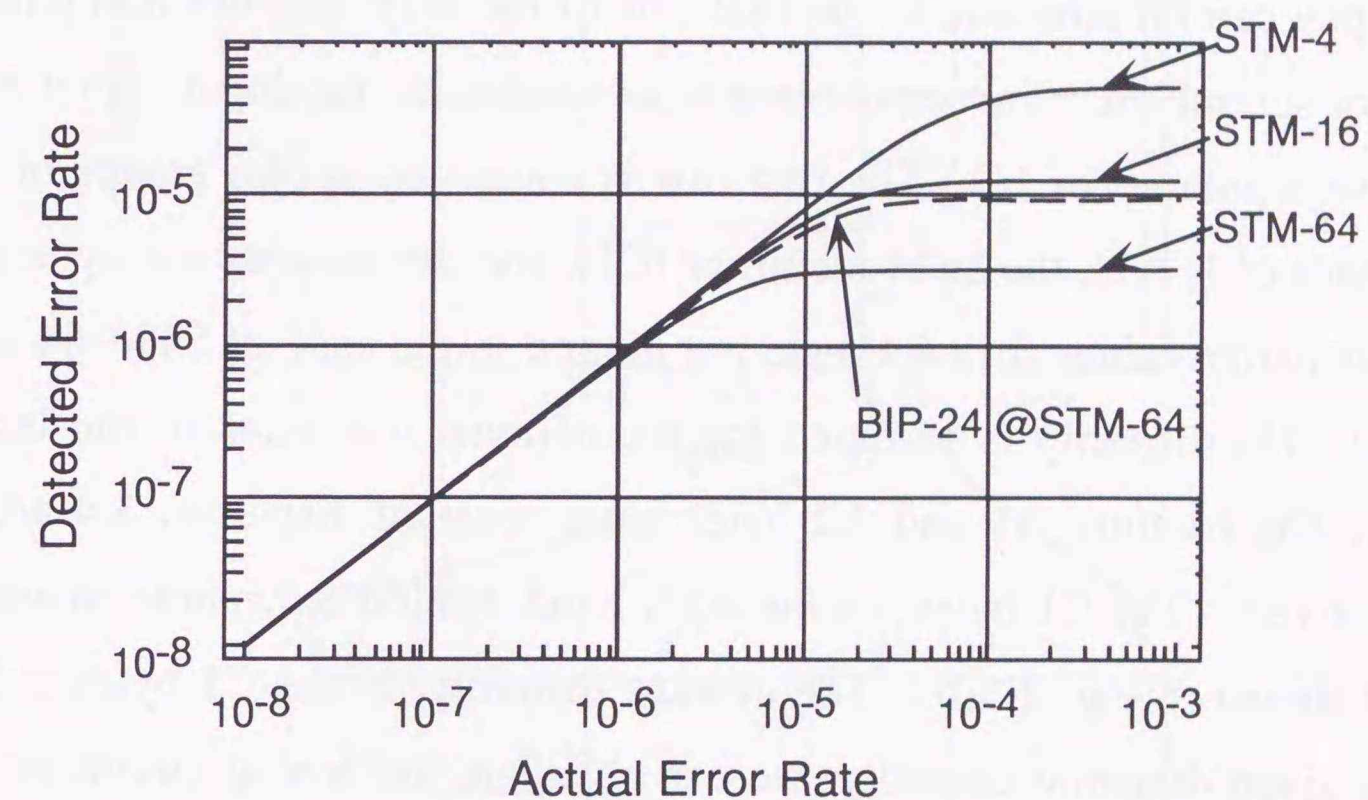


Fig. 2-4 BIP-8 accuracy of STM-4, -16, and -64. BIP-24 for STM-64 is also shown with dashed line.

should use BIP-8 or something else depends on how precisely we need to monitor the errors within each regenerator section.

2.3.3. Analog circuit requirements

All TDM systems are based on high-speed electronics, not only for digital circuits, but also for analog ones. For the STM-N frame in particular, we must pay careful attention to the first row of the SOH because it is polarized but not scrambled. The scrambling is to randomize the input signal and so achieve a polarity of $1/2$. The first row is composed of two groups of framing bytes (A1, A2), the STM identifier (C1), and the national use bytes (NU). The disparity values for each byte are plotted in the case of STM-64 in Fig. 2-5(a). No disparity is assumed for the national use bytes or the last 128 bytes. The framing A1 and A2 bytes have constant disparity, $3/4$ and $1/4$, respectively. The C1 bytes, on the other hand, exhibit a dynamic characteristic as shown in Fig. 2-5(b). The average disparity for the C1 bytes is 0.377. For a given dynamic disparity, we can calculate the analog circuit response needed. We assume a capacitor-coupled amplifier. Its output waveform dynamically displaces because the AC-coupled circuit can not sustain a direct voltage. Therefore, the signal level is displaced according to the input disparity. The waveform displacement was calculated and plotted in Fig. 2-6, for three low-frequency cut-offs; namely 1 MHz, 500 kHz, and 200 kHz. The peak-to-peak variations are about 30%, 13%, and 5%, respectively. Characteristic to the STM frame, the disparities continue over a specific time period. This is because the framing bytes are defined as a simple stack of $3N$ bytes for N multiplication, resulting in a constant period of unit byte length in terms of time. In other words, the problem becomes more serious as the multiplexing number increases. Other than this lower-frequency cut-off problem, we must also consider the response of the timing extracting circuit, which performs high-Q filtering and amplitude-limiting amplification. Since,

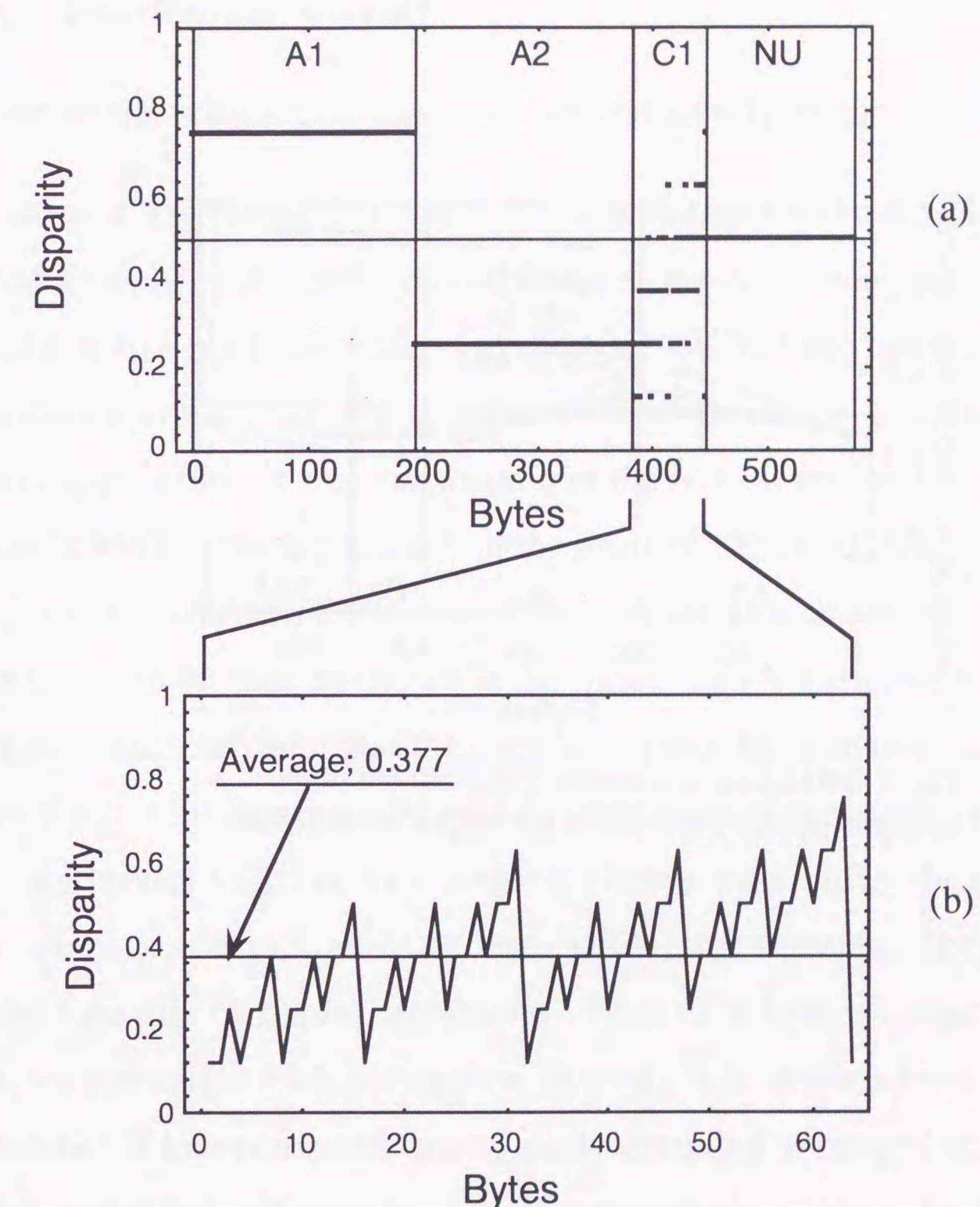


Fig. 2-5 (a) Disparity in the first row of the SOH

(b) Disparity in the C1 byte

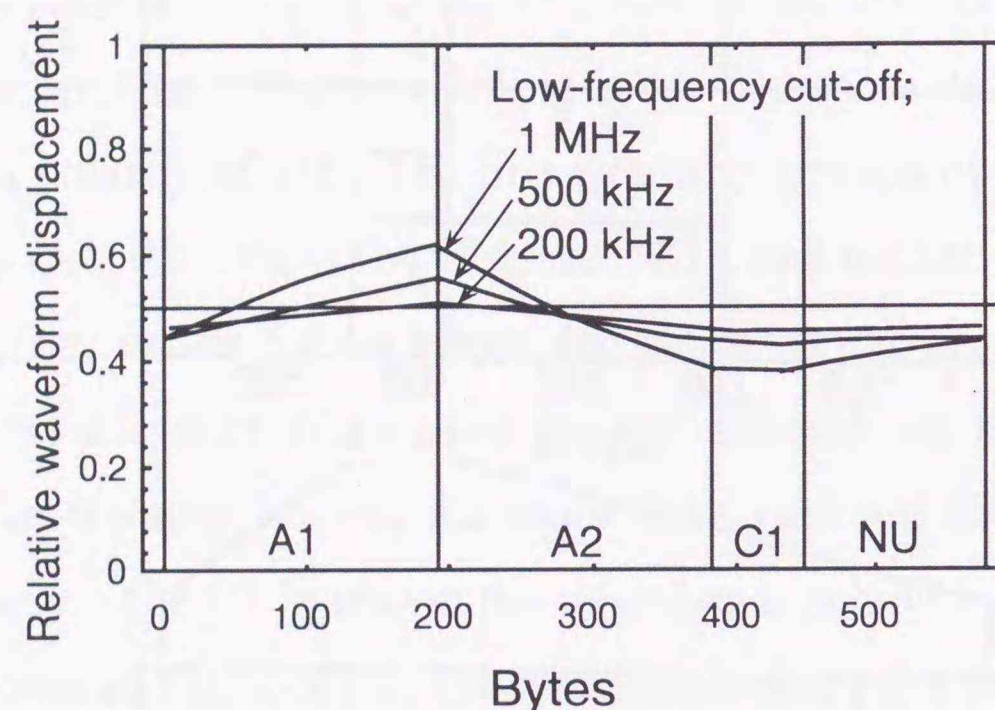


Fig. 2-6 Relative waveform displacement due to low-frequency cut-off in the amplifier response.

the response depends a lot on circuit detail, the response underwent further investigation and is reported in a letter [40]. In conclusion, the analog circuits must be able to accept dynamic disparities in the SOH bytes.

2.4. Byte Interleaving Circuit

2.4.1. Byte interleaving architecture --- General Configuration

To construct a large-capacity LTM within a reasonable and controllable scale, we propose a newly configured byte-interleaver. For N -channel multiplexing, it is composed of k ($= N/8$) row-column exchangers (RCXs) and one $N:1$ bit interleaver, as shown in Fig. 2-7. We introduced this basic idea in an earlier letter [41]. This paper optimizes the composition of the $N:1$ bit interleaver, which enables direct RSOH handing at 1/8th of the line rate. Each RCX has 8 input and 8 output ports, and converts 8 channels of byte-serial data into byte-parallel 8 outputs at the same bit rate, as shown in the figure. An $N:1$ bit interleaver combines all bits in such an order that the output sequence has a byte-serial format. To do this, the $N:1$ bit interleaver should be composed of two stages, namely eight $k:1$ interleavers followed by a single 8:1 interleaver. After the row-column exchange, we have k Most Significant Bits (MSBs), for example. They are then multiplexed timewise by the $k:1$ interleaver. Since each byte is composed of eight bits, we need eight such interleavers in total. It is obvious from Fig. 2-7 that the number of input channels can be easily extended in integral multiples of eight. This means byte interleaving circuit construction is simple for STM-8, 16, 24, 32, 40, and so on. Byte-interleaving from 156 Mbit/s to 1.24 Gbit/s was successfully achieved using a prototype circuit, as shown in Fig. 2-8(a) and (b). Between the two interleaving stages the STM- N signal is parallel, and this is where the SOH processing circuits are located. The multiplication ratio of the intermediate stage interleavers should be $2m:1$, because *odd number interleavers* (3:1 for example) are usually more difficult to implement at higher bit rates.

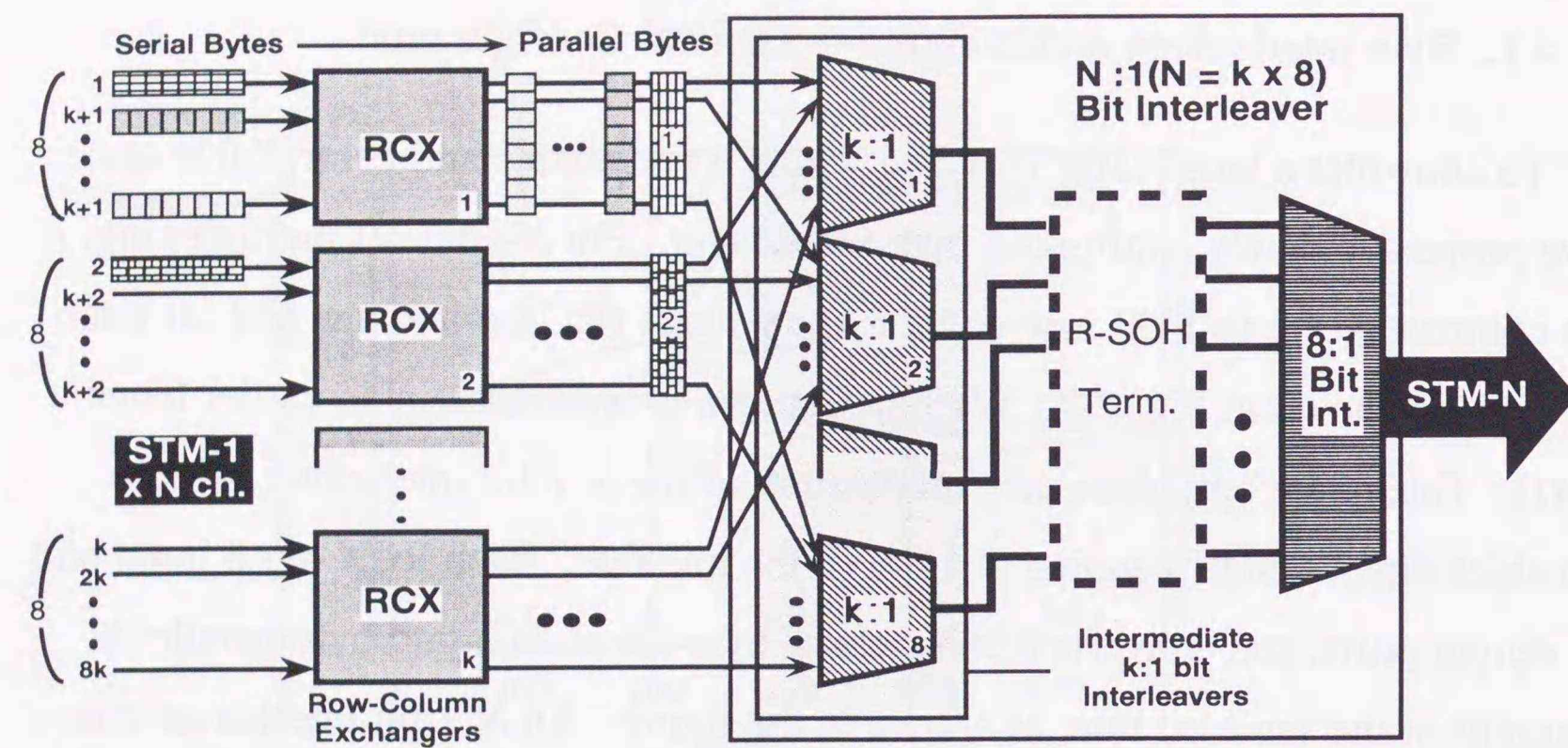
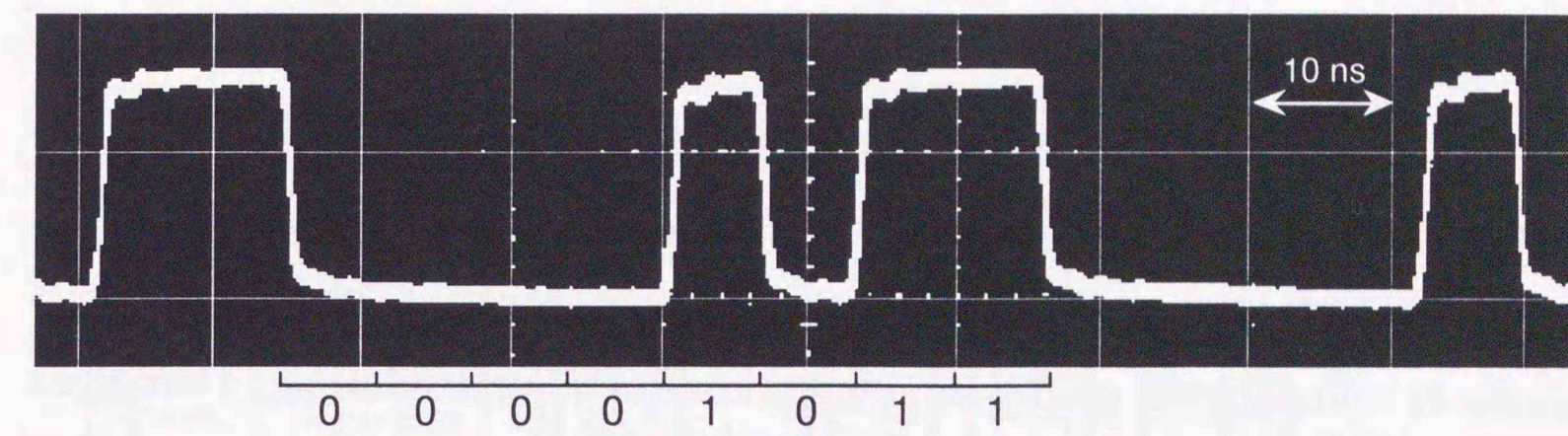
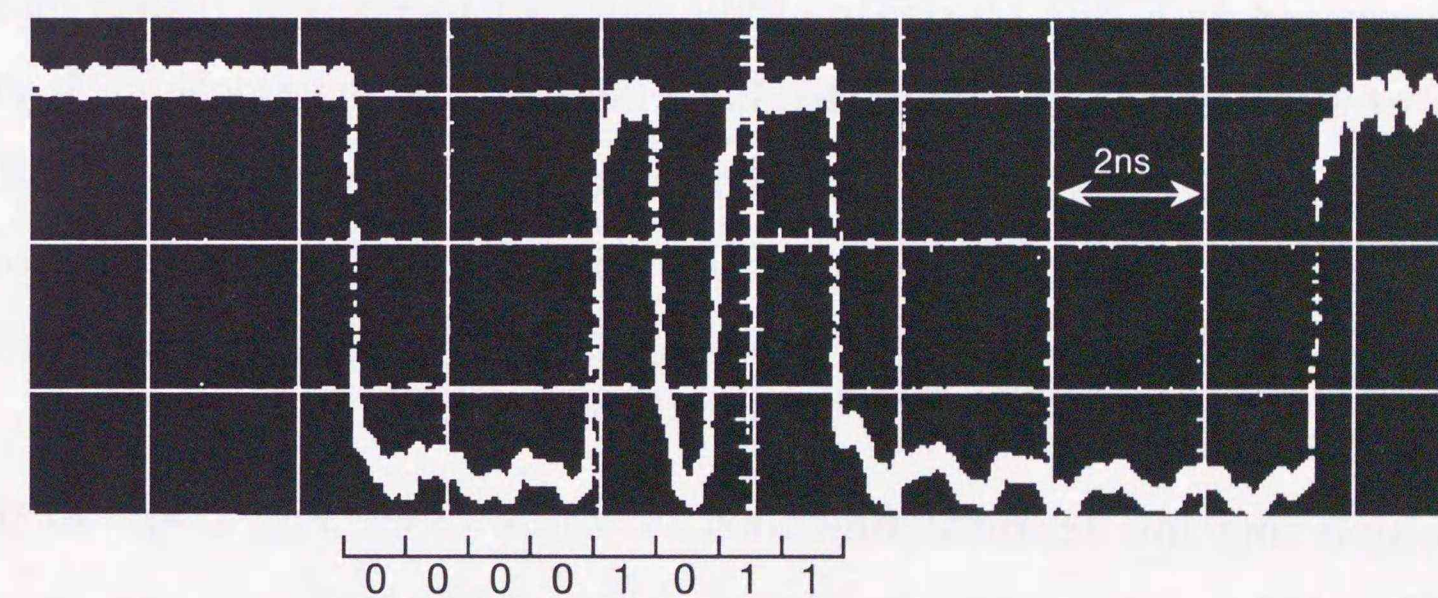


Fig. 2-7 N:1 byte interleaver. Each RCX has 8 input and 8 output ports, and converts 8 channels of byte-serial data into byte-parallel outputs at the same bit rate. An N:1 bit interleaver should be composed of eight of k:1 interleavers followed by a single 8:1 interleaver.



(a)



(b)

Fig. 2-8 Measured byte interleaver performance. (a) Experimental input signal at 156 Mbit/s. (b) Byte interleaved output at 1.24 Gbit/s.

With this consideration, the multiplexer configuration is very simple for $N = 8$ (1.25 Gbit/s), 16 (2.4 Gbit/s), 32 (5 Gbit/s), 64 (10 Gbit/s), 128 (20 Gbit/s), and so on. A prototype circuit for STM-64 byte interleaving was manufactured and tested successfully.

2.4.2. Prototype STM-64 multiplexing terminal

The implementation of an STM-64 line terminal using the byte-interleaving method is discussed below. We developed special ICs for the STM-64 LTM for two reasons: one is the multiplexed signal speed and the other is the enormous number of channels. Because signal connections between fast-switching circuits are extremely complicated, chip-wise integration is a natural solution. Two types of ICs, GaAs MESFET MSIs and Si Bipolar LSIs, were used to achieve the required operating speed and integration scale. Direct multiplexing from the fundamental bit rate (STM-1, 156 Mbit/s) enables the transmission cost to be reduced compared with step-by-step multiplexing. However, this results in an explosion of the number of lower channel circuits; 128 lower channels are needed for STM-64 LTM if a 1+1 protection scheme is employed for each of the lower channels. Thus, compact and low power consumption LSIs are important. We fabricated a utility LSI with Bi-CMOS technology to realize an LTM with reasonable circuit scale.

The multiplexing line terminal functions as follows (see Fig. 2-9). In the figure, each bold boxes represents developed prototype IC chip.

- (i) At the incoming stage of the multiplexer, each low-bit-rate channel is terminated. Processing the AU pointer aligns the frame phase of the 64 incoming channels to the reference phase. For each channel, such transmitter and receiver functions are performed by a Bi-CMOS LSI with about 50 kgates. Even for STM-L ($L < N$) inputs, their multiplex section overhead (MSOH) has to be processed on an STM-1 basis, but the pointer must be adjusted simultaneously for the STM-L bulk signal. Therefore, an LSI

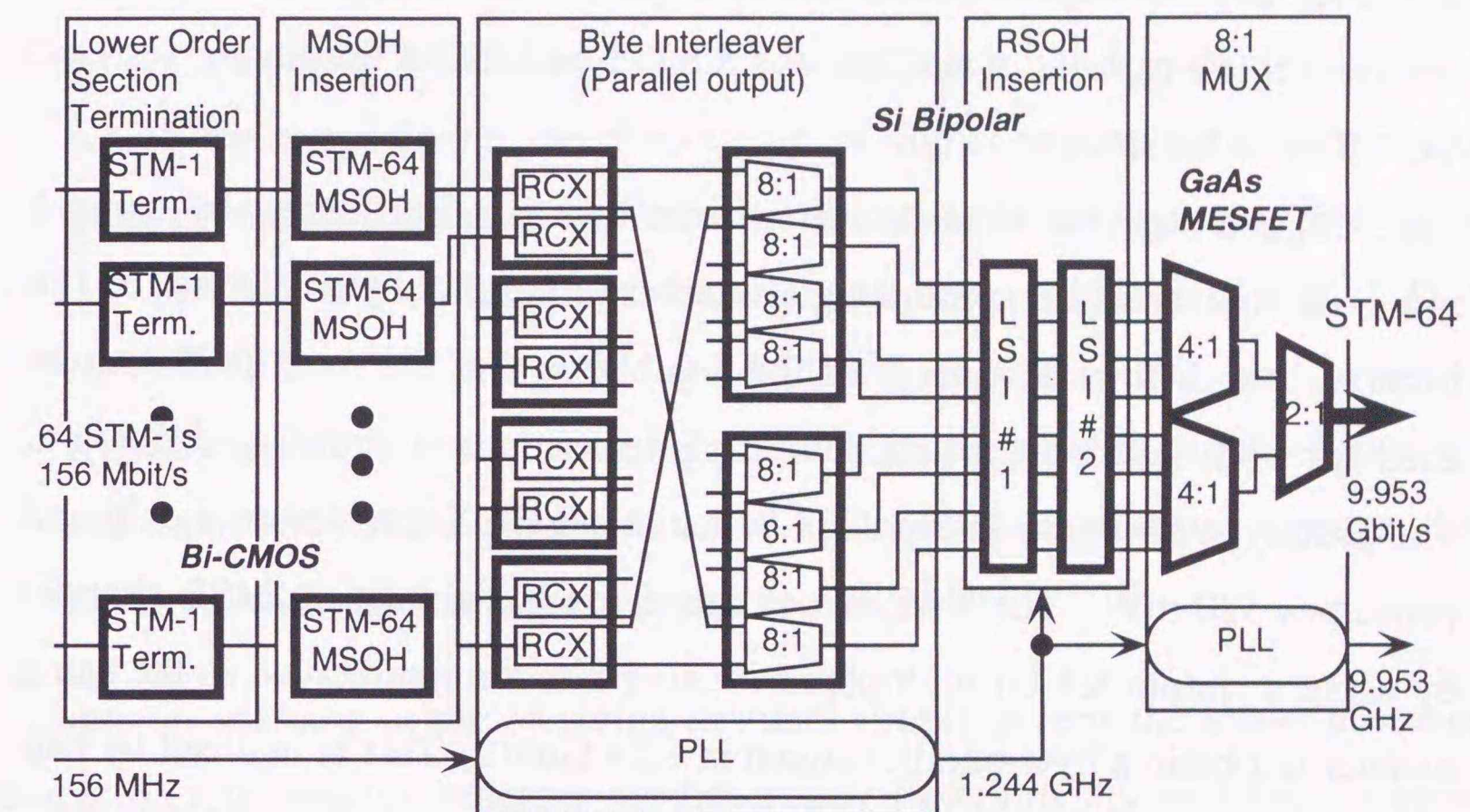


Fig. 2-9 An STM-64 line terminating multiplexer configuration. Each bold box represents developed prototype IC chip.

supporting the STM-1 process and pointer-concatenation is sufficient for any multiplexed input STM-L. An additional necessity is a byte interleaver between STM-L and corresponding STM-1 channels. The LSI has a low power consumption feature of only 1 W, which is 20% of the power consumed by an equivalent discrete circuit.

- (ii) 64 identical Bi-CMOS LSIs construct the MSOH for the STM-64 frame. BIP-24N is processed at this level. The Bi-CMOS LSI described in (i) is designed to be capable of serving as an STM-64 MSOH terminator for each VC-32 or -4 tributary.
- (iii) The aligned signal is then converted into byte-parallel format by using 8 row-column exchangers, producing 8 bundles of parallel bytes clocked at 156 Mbit/s. The distinct advantage of this method is that the same RCX can be used for both multiplexer and demultiplexer circuits without modification. The prototype 900-gate Si-bipolar LSI mounts 2 RCX circuits on a chip and consumes 520 mW. The 8 MSBs are interleaved into a single MSB channel by using a simple 8:1 bit multiplexer. Other bits are interleaved in the same manner to obtain a byte-parallel signal at 1.24 Gbit/s. This is realized by two LSIs accommodating four 8:1 bit multiplexers each consuming 2 W. The inverse process at the receiving side is performed by four LSIs having two 1:8 bit demultiplexers each consuming 2.5 W. The Si-bipolar process (SST) was used to allow the interleavers to operate at 1.24 Gbit/s. Multiplexer ICs are furnished with clock dividers to generate lower clocks from higher ones. This requires a clock pump-up function to raise the input 156 MHz to the multiplexing speed of 1.24 GHz. A Phase-Locked Loop (PLL) is used as shown in Fig. 2-9, so that once the loop is locked, the phase difference between input and generated clocks is kept constant. By using an elastic memory, the independency of the two clocks can be absorbed at the cost of increased signal delay between writing and reading.

- (iv) Transmitter and receiver RSOH processing functions are divided between two Si-bipolar gate arrays because of the speed-scale trade-off. Each Si-bipolar gate array LSI accommodates 4000 gates. Framing, BIP-8 counting, and scrambling are performed by the transmitter LSIs consuming 10.4 W in total. Similar size chips consuming 9.5 W in total are used for frame detection, BIP-8 verification, and descrambling at the receiving side.
- (v) The byte-parallel 1.24 Gbit/s signal is multiplexed to yield 10 Gbit/s (9.95328 Gbit/s precisely) serial data. The 8:1 multiplexer and demultiplexer, both of which are very simple, need no complex signal processing capability, but must have switching rates not less than the transmission line rate. We utilized 2:1 and 4:1 bit multiplexers and demultiplexer MSIs realized as GaAs MESFETs with 0.2 μm gate length [42, 43]. A sample 10 Gbit/s byte interleaved waveform is shown in Fig. 2-10 with the 10 GHz clock. Since only one input 156 Mbit/s channel is active and conveys a random sequence, the 8 bit block with random pattern appears to be isolated.

Signal processing at the receiving terminal simply inverts the above processes. Since the regenerative repeater performs only functions #iv and #v, the circuit scale can be smaller than that expected from bit-interleaving systems with equivalent functions.

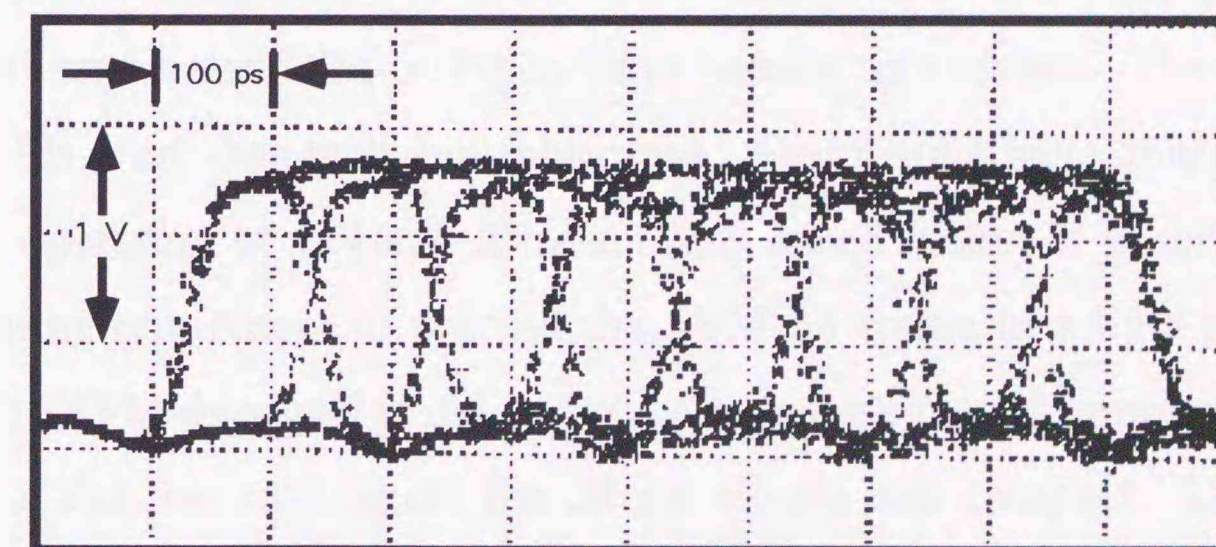


Fig. 2-10 Byte interleaved 10 Gbit/s output waveform.

2.5. Summary

By comparing WDM, and two TDM schemes (bit interleaving and byte interleaving), we conclude that TDM with byte interleaving is best suited for the near future SDH network system. We have proposed an architecture for a line-terminating multiplexer (LTM) that realizes simple circuit configuration by multiplexing channels in integral multiples of eight. The key circuit is 8-channel row-column exchanger (RCX), which we have implemented as a Si-bipolar LSI. We have also shown that the STM-64 LTM can be successfully constructed with Bi-CMOS, Si-bipolar, and GaAs MESFET ICs. Test results show that an STM-64 or 10 Gbit/s line-terminating multiplexer is feasible.

Chapter III

High-speed Regenerator Section Terminating LSIs

3.1 Introduction

Research and development activities are being undertaken for the next generation of large capacity transmission systems. The author has proposed the STM-64 [44], which is capable of 9953.28 Mbit/s transmission, as a simple four-fold extrapolation of the STM-16 (2.5 Gbit/s) architecture. This chapter reports a four-chip LSI set for the STM-64 regenerator-section termination. The Si-bipolar gate-array LSIs were designed and fabricated. Test results have confirmed their successful operation at 1244.16 Mbit/s. This speed makes it possible for the line-terminating equipment to process the STM-64 frame in a byte-parallel manner. Since the STM-64 signal is defined to be byte multiplexed, processing clock speed of 1/8 of the line rate makes the circuit simple and compact. Lower clock rate relaxes restriction in design and manufacturing process, however, it increases circuit complexity and its scale.

3.2 Basic Requirements

Each line terminating multiplexer (LTM) and regenerative repeater (R-REP) must be equipped with a regenerator-section terminating circuit that hosts the following functions:

- Frame alignment <A1, A2>,
- Bit interleaved parity (BIP) calculation and verification <B1>,
- Fault location <F1>,
- Order-wiring <E1>,
- Data Communications Channel (DCC) bytes drop and insertion <D1 - D3>,

where byte names assigned in the STM frame are bracketed. Rigorously speaking, the frame alignment function is excluded because it is defined as a part of the SDH Physical Interface function (SPI) by G.783 recommendation. Since no serial framing circuit can support 9.96 Gbit/s, it is a natural consequence that a parallel framer circuit is integrated into the regenerator section termination function (G. 783) block, which processes the regenerator section overhead (RSOH) bytes. Byte-based frame indicators (A1, A2) are best handled in a byte-parallel manner. Attention has to be paid to framing with the byte-parallel format because there is no way of knowing when a byte begins or byte phase at the 9953.28 Mbit/s serial sequence before the frame is aligned. Two methods are compared. The first is to reset the demultiplexer or serial-to-parallel converter when the byte phase is detected by the framing circuit. This detection process will take one frame period in the worst case. The actual framing process is started thereafter, and may also take another frame length. The overall framing period for the worst case will be equal to two frames period (250 μ s), which is the maximum frame alignment time for an error-free signal with no emulated framing patterns [45]. The framing time can be reduced if framing circuit interchanges the parallel inputs to align the bytes. Accordingly, the receiver LSI is equipped with a byte phase alignment circuit in the front end, as shown in Fig. 3-1-b. The byte-parallel nature of the inputs leads

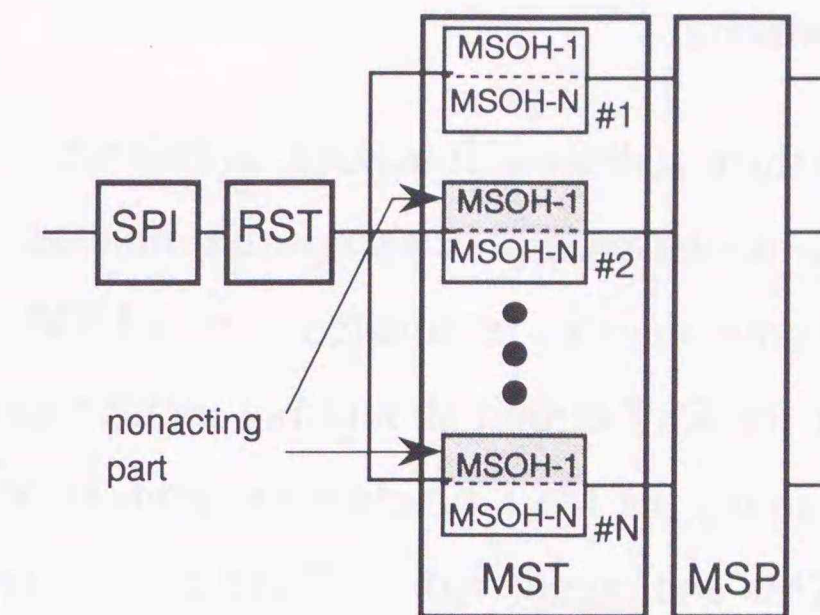


Fig. 3-1 (a) Functional block in rigorous accordance with the ITU recommendation (G. 782)

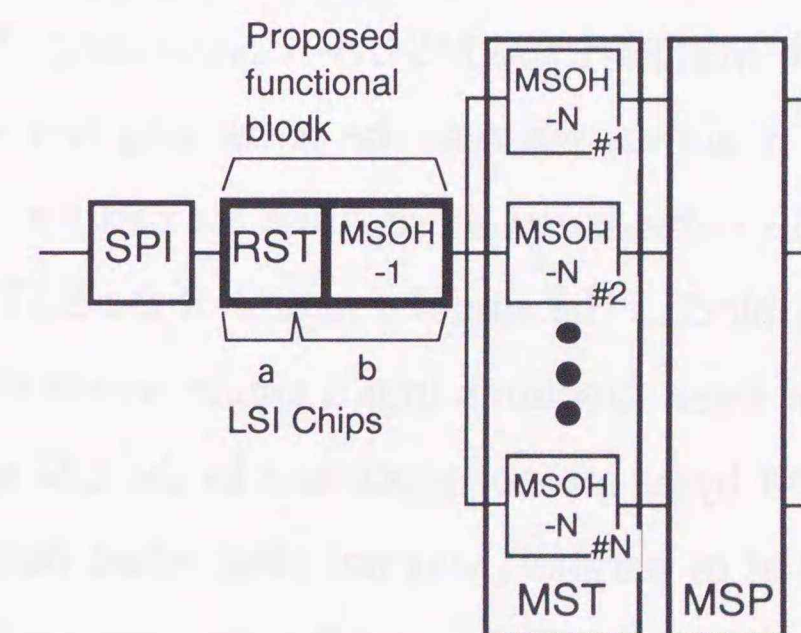


Fig. 3-1 (b) Proposed functional block

to a processing speed of one-eighth the line rate, or 1244.16 MHz at which Si-bipolar LSIs are operational.

3.3 Additional Requirements

The LSI set features three additional functions, as follows.

The first is one part of the multiplex section termination (MST) function. Because the multiplex section must be terminated at each LTM, not regenerators, it seems reasonable that the RST circuit should not include any MST function. However, we should examine the MST function to optimize overall circuit size considering both the LTMs and regenerators. There are two types of multiplex section overhead (MSOH) bytes; the first is defined only in a single set for each STM-N frame, such as K1, K2, E2, and DCCM bytes. The other type includes B2, Z1 and Z2, which are defined in N sets for the STM-N multiplex frame [46]. We call the former MSOH-1 bytes and the latter the second MSOH-N bytes. If we construct the LTM circuit in rigorous accordance with the function block defined by ITU, the circuit should be redundant. It has N circuit blocks being capable of processing both MSOH-1 and MSOH-N processing. However, only one of the MSOH-1 circuit is active, whereas the remaining N-1 circuits must be kept inactive. To avoid this redundancy we separate the part for the MSOH-1 from the MSOH-N processing block. The singular nature of the RST and MSOH-1 circuit reasonably combines these functions into a single terminating LSI set. Despite this argument, DCCM bytes are not processed in the LSI set, because the detail handling format (serial or parallel) was not clear when designed. The MSOH-1 function must be idled if the LSIs is used for the regenerators. When rewriting the MSOH-1 bytes, the transmitter LSI also rewrites the corresponding B2 bit(s) accordingly to avoid a mismatch in the BIP-N x 24.

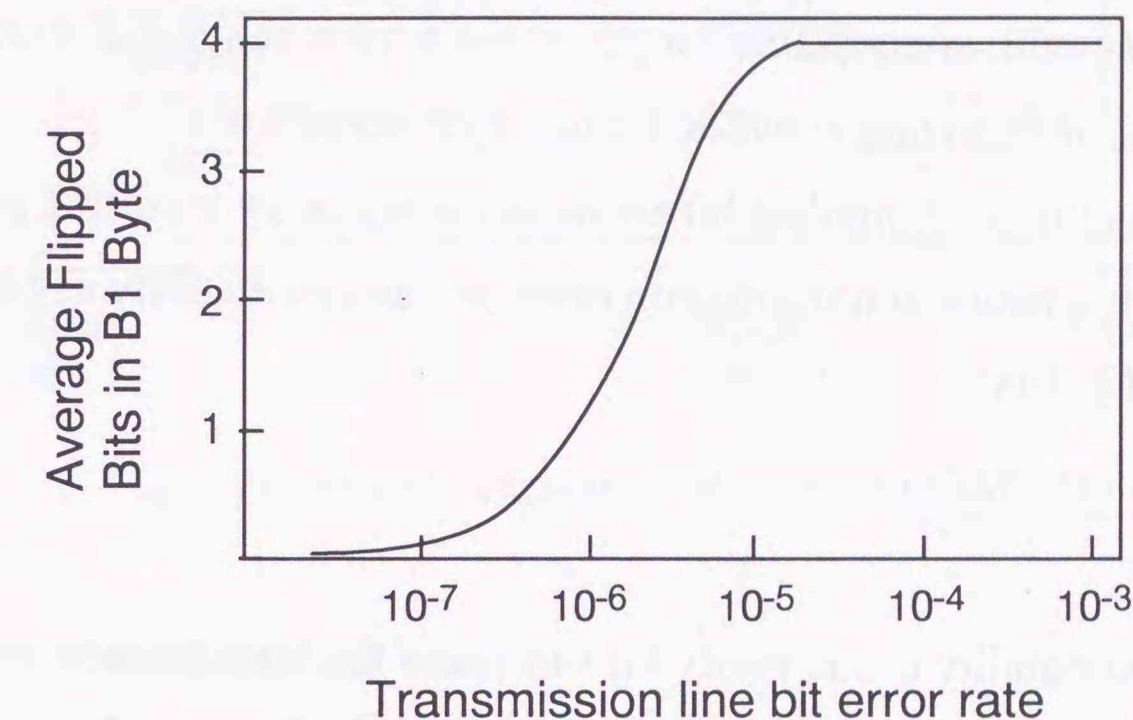


Fig. 3-2 Average number of B1 bits flipped by BIP-8 error detection.

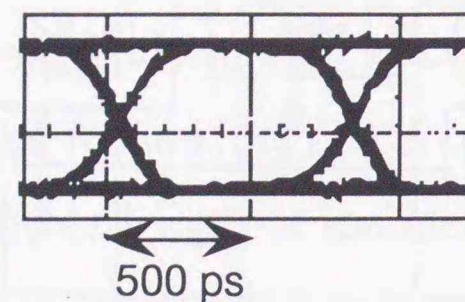


Fig. 3-4 Output waveform of ECL interface of transmitter LSI-a at 1244.16 Mbit/s

Table 3-1: Estimation of gate volume and measured power consumption

LSI	Gates used	Power consumption
	kgate	W
Tx-a	2.9	6.3
Tx-b	2.7	6.8
Rx-a	2.2	5.5
Rx-b	2.7	5.7

3.4 Prototype LSIs

A prototype LSI set was manufactured on SST-1B gate arrays [49]. Two gate-array LSIs each having 4 k gates of Si-bipolar transistors, were used to develop the transmitter and receiver functions respectively. Those four LSIs were separately mounted on 120-pin flat packages. The Si-bipolar gate-array LSIs have ECL inputs and outputs interfaces and their successful operations at 1244 Mbit/s were confirmed. Fig. 3-4 shows a typical waveform from LSI-a with showing the good eye opening achieved. An estimation of the gate volume and measured power consumption is listed in Table 3-1.

3.5 Summary

A four part regenerator section termination LSI chip set has been designed, fabricated, and successfully tested. Its additional functions include frame alignment, some multiplex section termination functions, test pattern generation, and virtual error insertion. The Si-bipolar gate-array LSIs process the STM-64 signal in the byte (8 bit) parallel manner at 1.24 GHz. This result confirms feasibility of the line-terminating equipment for the STM-64 system.

Chapter IV

Supervisory Signal Transmission by Modulating the mB1C Code Complementary Bit

4.1 Introduction

Supervisory subsystems provide in-service monitoring and fault location for dependent repeaters. The use of metallic conductors, separate fibers, or separate system is impractical in optical fiber transmission systems. It is desirable to develop a supervisory system which transmits both traffic and supervisory information over the same optical transmission line simultaneously. Thus far three proposals applicable to a digital system have been examined.

- (a) Supervisory pulse insertion: Sampled supervisory digital pulses are inserted between shortened traffic pulses. This results in transmission line clock frequency increase, undesirable in high-speed transmission systems. It also requires repeaters to reconstruct line code sequences to transmit operation

information to terminals. So a sophisticated and large-scale repeater circuit is necessary.

- (b) System clock frequency modulation: The system clock frequency is slightly modulated by the supervisory information, which passes through the other repeaters if their timing recovery circuits have wide enough bandwidths. This technique requires no regeneration of the supervisory signal at repeaters [50]. Peak clock phase deviation must be small to avoid impairing the main data. In addition, this is unsuitable for sending signals from terminals to repeaters in land systems, because repeaters must be uniquely identifiable for interrogation purposes. Unlike submarine systems, it is more practical for land repeaters to be noncustomized.
- (c) Frequency division multiplex, or inserting a supervisory data channel below the traffic band: This technique requires a low disparity code such as mBnB for traffic data to decrease traffic and supervisory spectrum overlap [51]. It is very difficult to construct an mBnB code converter at high bit rate (>0.5 Gbit/s) on a reasonable scale. Thus this technique is unsuitable for gigabit systems.

This chapter presents a novel transmission method for a supervisory digital information applicable to gigabit systems over the same fiber. It minimizes circuit scale, clock frequency increase, and interference between traffic and supervisory information. And it does not require customization.

4.2 Modulation of Complementary Bits in mB1C Code

The mB1C code has been adopted in F-400M and F-1.6G optical trunk transmission systems [52], because it needs less circuits than does mBnB code [53]. The code adds a single bit after every m bit block as a complement to the last bit of the block. Clock frequency increase is kept at a level of $(m+1)/m$. This additional bit is called "C bit." If we modulate the C bit's amplitude

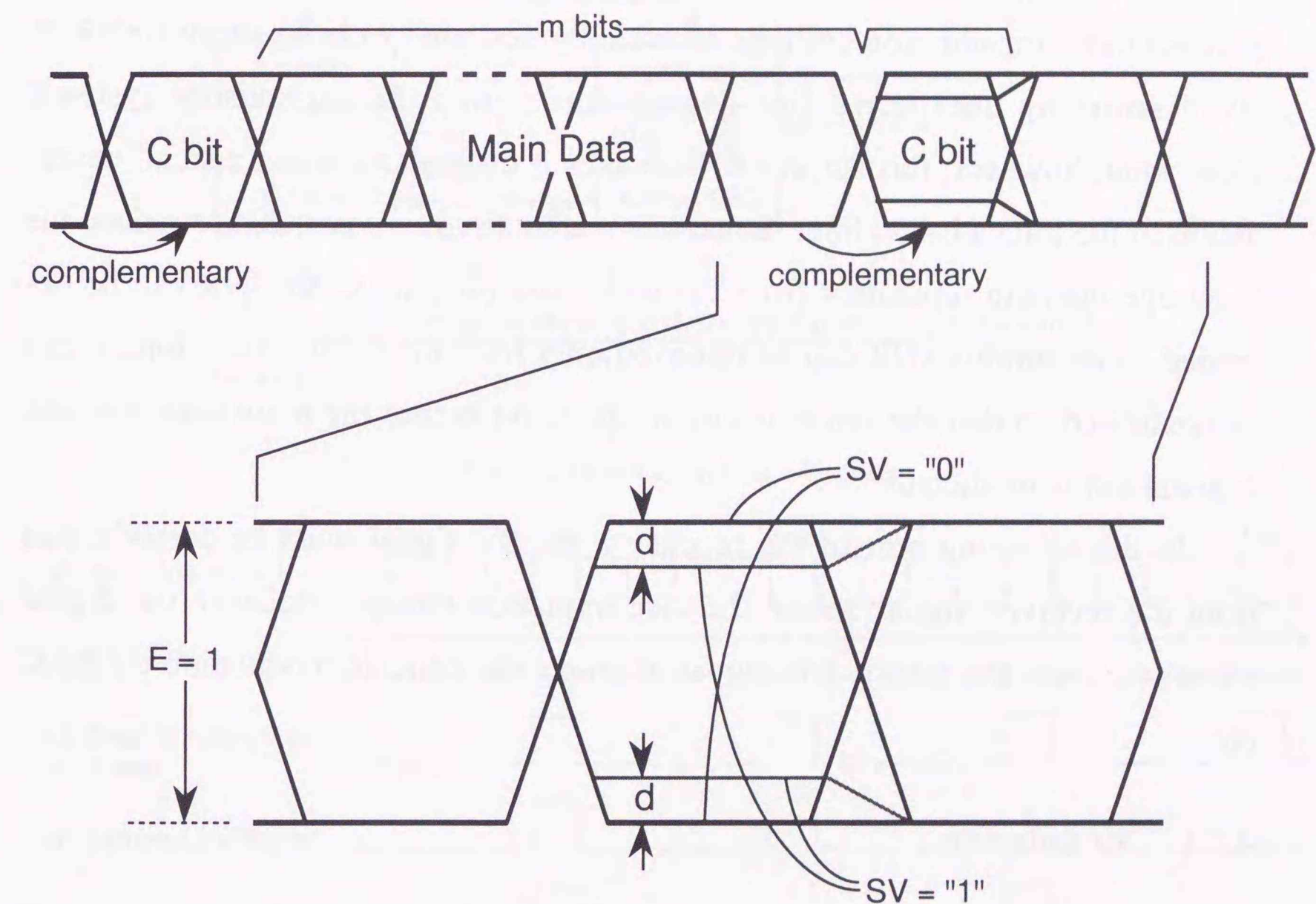


Fig. 4-1 Eye diagram for C bit modulation

according to the supervisory (SV) signal and do not modulate the main data block, we can isolate traffic and supervisory information in time over the same fiber transmission.

An additional advantage is that the transmission line degradation, such as fiber loss increase and/or transmitter light intensity decrease, can be detected before it impairs the traffic information. In order to ensure "error free" transmission, the data bits must have sufficient signal-to-noise ratio (SNR). Furthermore, transmission systems are usually designed to have surplus SNR to avoid errors by unexpected line degradation. From the supervisory systems' view point, however, this surplus SNR makes it difficult to detect a small degradation in the transmission line. Since this method makes it possible to reduce the C bit eye opening separately from the main data bits, the C bit SNR can be degraded. The surplus SNR can be removed only from the C bit. Therefore, it can be controlled so that the errors occur on the C bit before the main data; the line degradation is predictable.

In the receiving port of the repeaters, the SV signal must be demodulated from the received signal before the discrimination circuit. Because the digital circuit recovers the binary data signal, it erases the delicate modulation on the C bit.

4.2.1 Modulation

The C bit is complementary to its preceding bit. An eye diagram for the C bit modulation is shown in Fig. 4-1. Here, we assume a binary SV signal. In principle, the SV signal is not necessarily digital, but can be analog. When the SV signal is low (0), the C bit's eye opening (E) is kept the same as that of the other bits; $E=1$. When the SV is high (1), the eye opening is reduced to $E=1-2d$, by subtracting " d " from the C bit pulse and by adding a " d " amplitude pulse to the absent C bit time slot. A modulation block diagram and corresponding timing diagrams are shown in Figs. 4-2 and -3, where a 4B1C code is shown

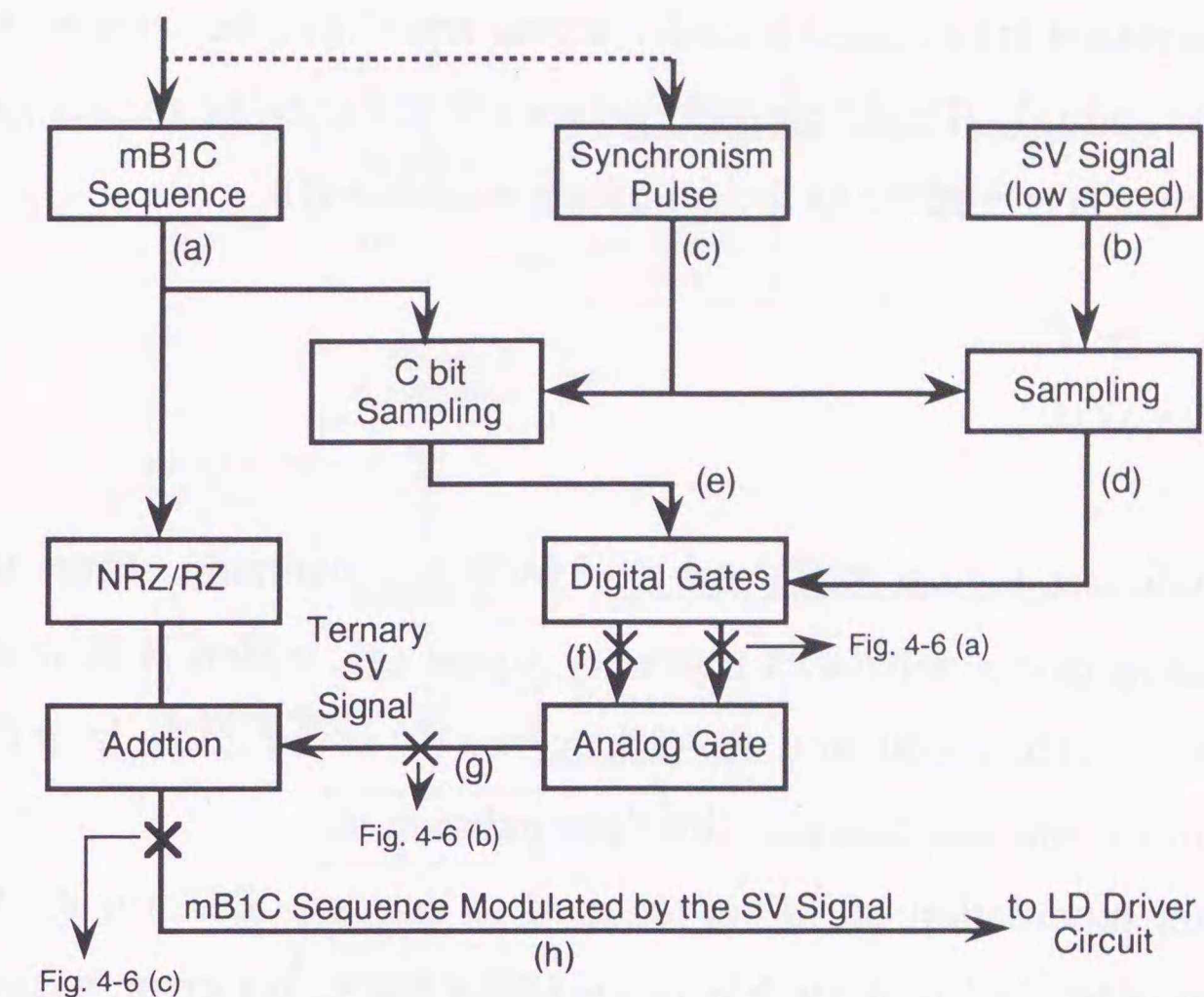


Fig. 4-2 Modulation block diagram

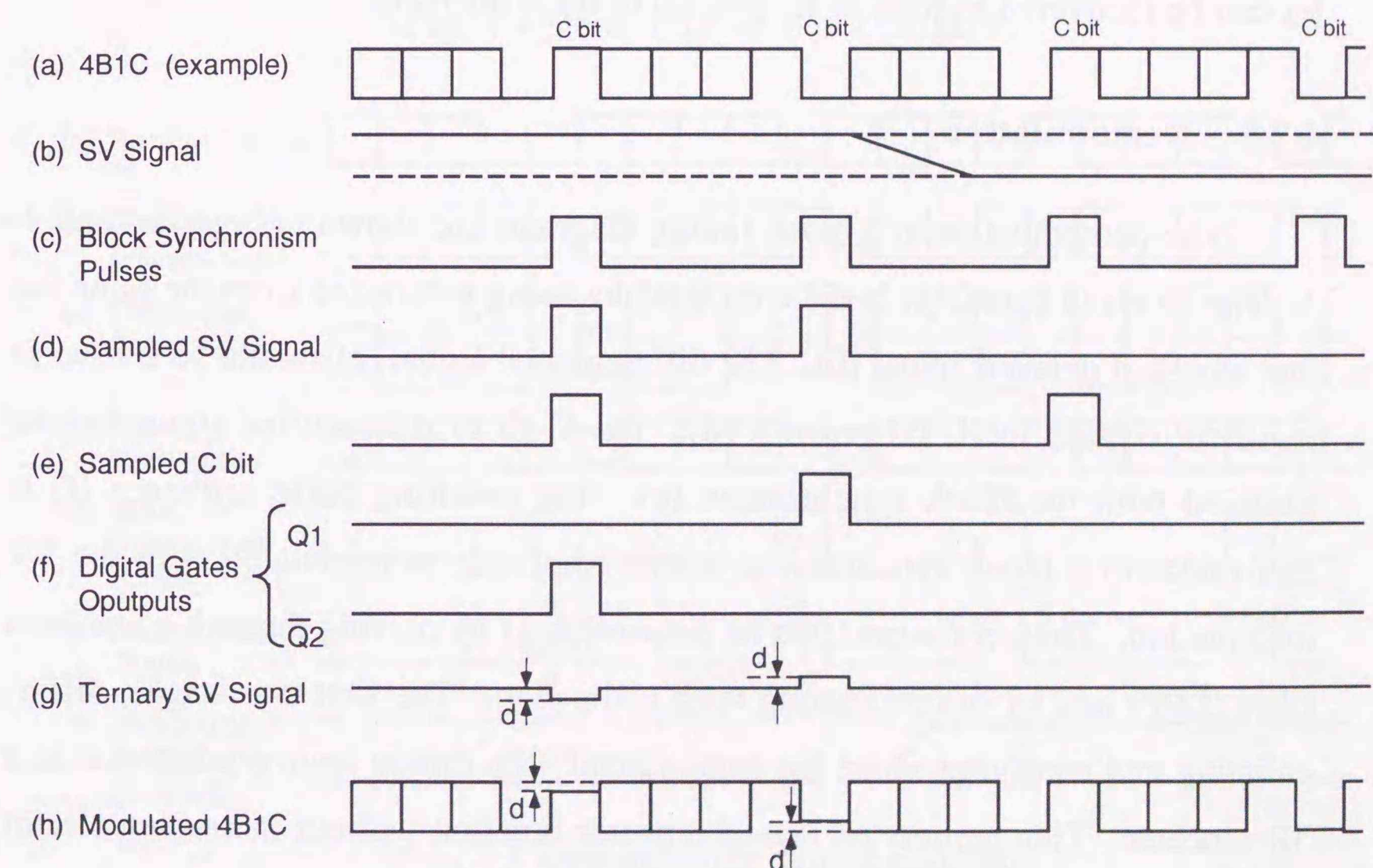


Fig. 4-3 Modulation timing diagram

as example (a). The SV signal is assumed to be low bit rate (b). The block synchronism pulses (c) samples both SV signal and C bit, the sampled signals are shown in (d) and (e). These sampled pulses are fed into the digital gates to get Q_1 and Q_2 signals according to the following relations (f):

$$Q_1 = SV \cdot \bar{C}$$

$$\bar{Q}_2 = SV \cdot C$$

where " $\bullet$ " indicates logical AND and " $\bar{}$ " indicates inversion. From those signals, an analog gate synthesizes a ternary signal (g), which is to modulate the 4B1C sequence. The modulated 4B1C sequence (h) is fed into a laser diode (LD) driver circuit to generate transmission light pulses.

Greater modulation of " d " is required to increase SNR for the SV signal. However, smaller " d " is desirable to suppress the C bit error because of the reduced eye opening when $SV = "1."$ It is important to point out that the erred C bit can be recovered as long as its preceding bit is unerring.

4.2.2 Demodulation

The demodulation and block timing diagrams are shown in Figs. 4-4 and 4-5. The received signal (a) is differentiated by being subtracted from the same but one time-slot delayed signal (b). The differentiated signal (c) should be full-wave rectified to erase the C bit polarity (d). The C bit of the rectified signal is now sampled with the block synchronism (e). The resulting pulse sequence (f) is equivalent to a block synchronism whose amplitude is modulated with the SV information. Thus, SV signal can be recovered (g) by passing through a low-pass filter (LPF) and by discriminating with a flip-flop. The first two stages, differentiating and rectifying, share the same circuit with timing recovery function in a 3R repeater. This permits increased repeater function without increasing circuit size, a distinct practical advantage.

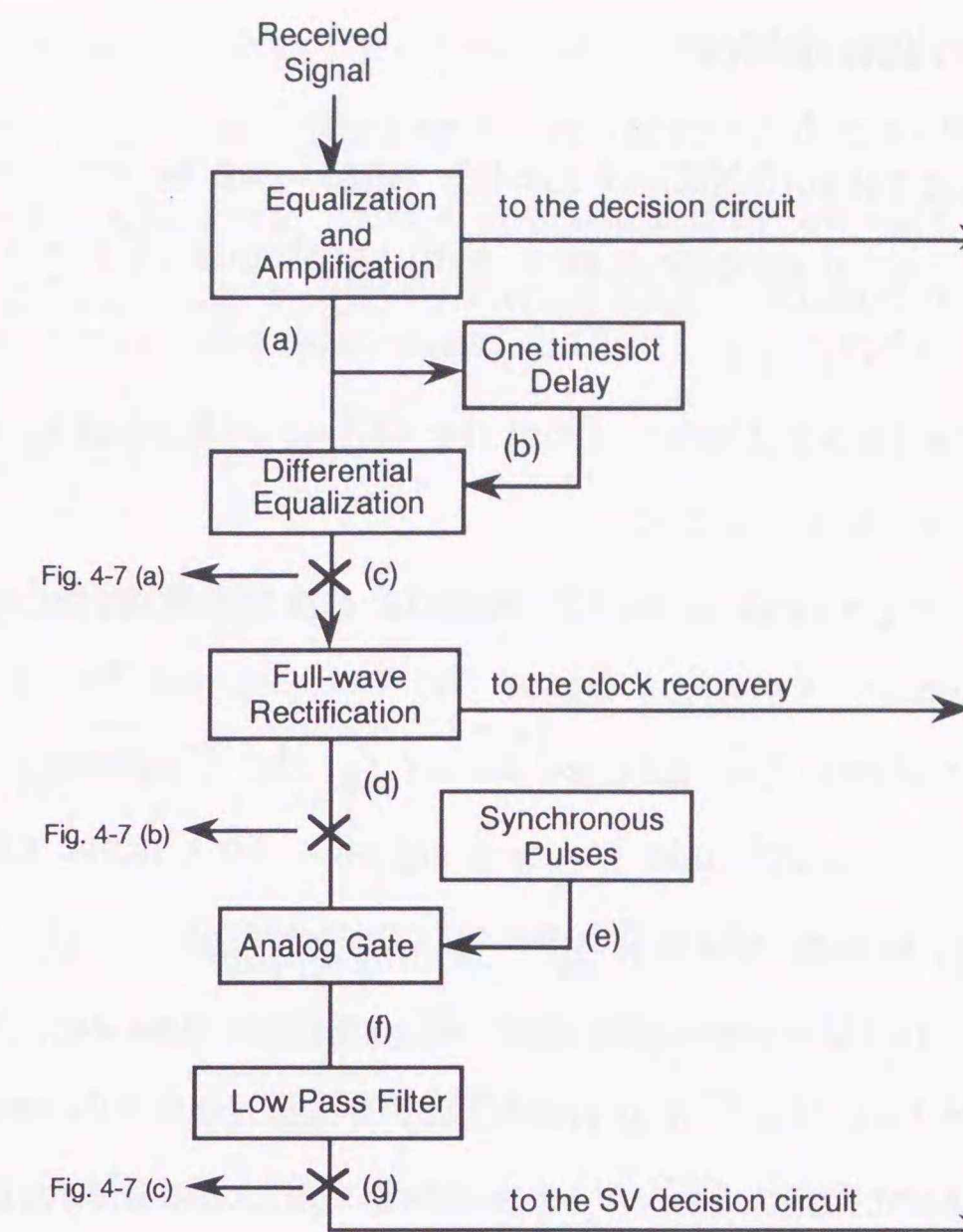


Fig. 4-4 Demodulation block diagram

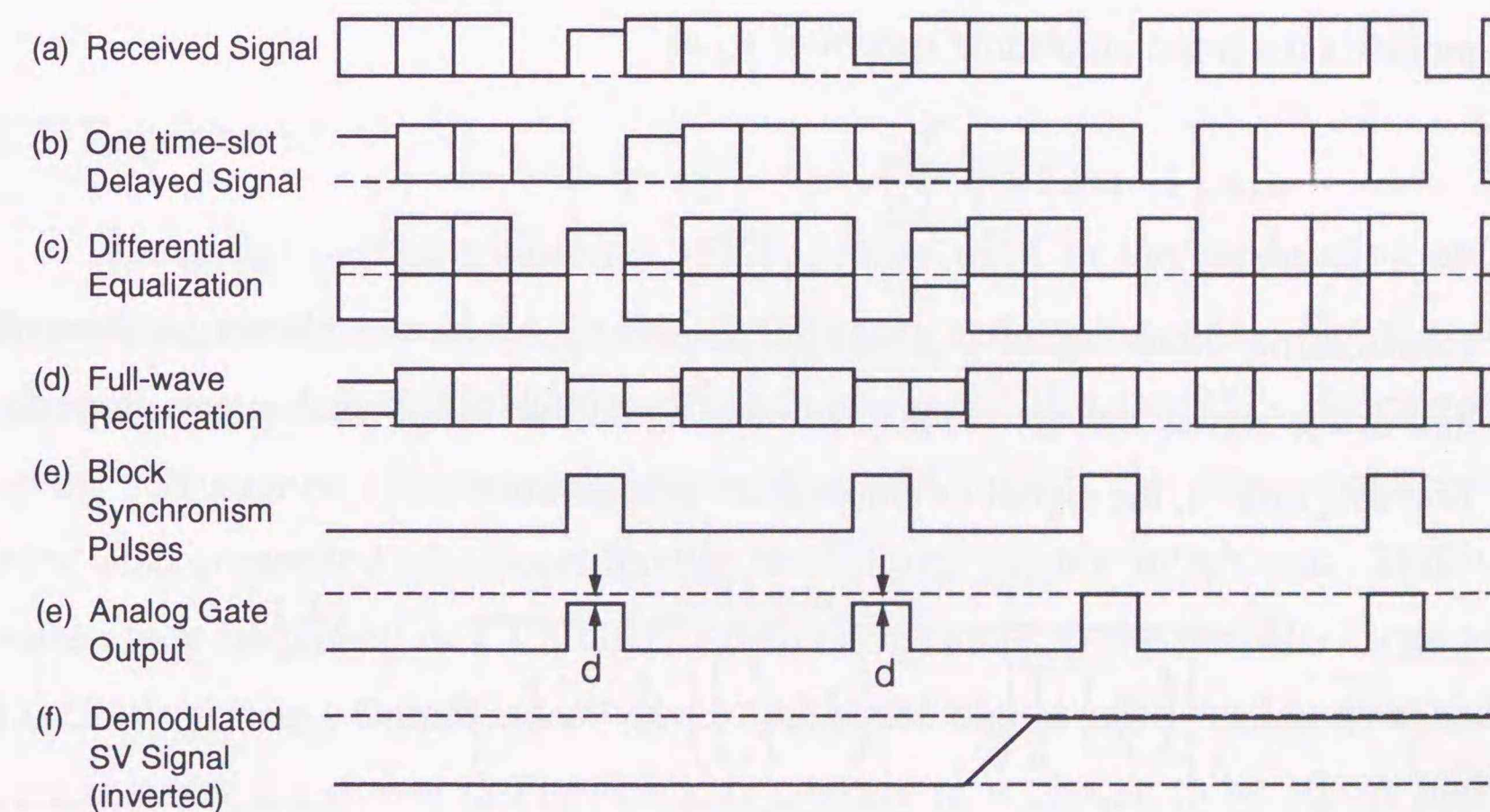


Fig. 4-5 Demodulation timing diagram

4.2.3 Noise Characteristics

Signal-to-noise ratio (SNR) of the SV signal can be estimated as follows. The normalized SV signal spectra appear with amplitude $d[2(m+1)]$ at frequencies $nf_0/(m+1)$, where f_0 is the main-data line clock frequency and $n = 0, 1, 2, \dots$, and the factor 2 arises from the differential equalization process, in which the signal is divided into two.

The subtraction process in the differential equalization reduces the random noise around frequency $f = nf_0$. Since the two signals being subtracted are originally the same signal but delayed by $1/f_0$, the frequency components at $f = nf_0$ have the same amplitude in each signal. This noise can be canceled. However, inversely, intermediate frequency components $f = [(n+1)/2]f_0$ have the same amplitude but the opposite sign. The subtraction enhances these noise components. Therefore, very low frequency noise ($n \approx 0$) can be greatly reduced with the subtraction. Here, we assume the normalized noise amplitude before the differential equalization with $i_m(\text{Hz}^{-1/2})$. We put the main data $\text{SNR} = N_m(\text{dB})$, which gives $N_m \approx -10 \log(i_m^2 f_0)$. The differential equalization modifies the noise amplitude spectrum n_d as:

$$n_d = \left(1 - \cos 2\pi \frac{f}{f_0}\right) \frac{i_m}{2}.$$

Considering other signal degradation factors, such as the circuit nonlinearity and/or the signal leakage from main data signal due to the analog gate imperfection as $i_a(\text{Hz}^{-1/2})$, the signal to noise ratio Q is given as

$$Q = \frac{d^2/(m+1)^2}{\left[\left(\frac{i_m}{2}\right)^2 \int_0^{f_{sv}} \left(1 - \cos\left(2\pi \frac{f}{f_0}\right)\right)^2 dt + i_a^2 f_{sv}\right] n_f} \quad (4-1)$$

where the filter response is approximated to be rectangular of f_{sv} width and a noise increase factor n_f due to the SV signal amplifier. The first term of the

denominator represents the noise reduction by the differential equalization. The second term represents the effective noise increase due to the circuit imperfections. Because this chapter's interest is restricted to the very low frequency relative to the clock frequency f_0 , the following approximation is acceptable:

$$1 - \cos\left(2\pi \frac{f}{f_0}\right) \approx \frac{1}{2} \left(2\pi \frac{f}{f_0}\right)^2.$$

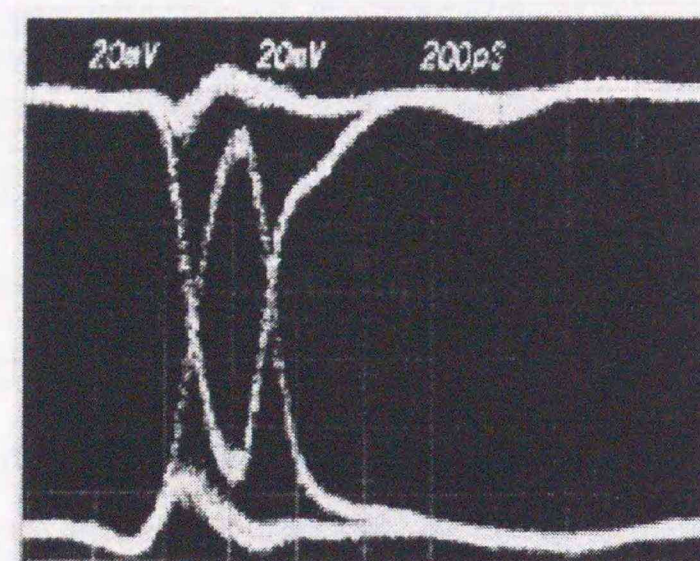
Thus the SV signal SNR (Q) can be approximated as

$$Q \approx \frac{d^2}{4(m+1)^2 \left[\frac{\pi^4 i_m^2 f_{sv}^5}{5 f_0^4} + i_a^2 f_{sv} \right] n_f}. \quad (4-2)$$

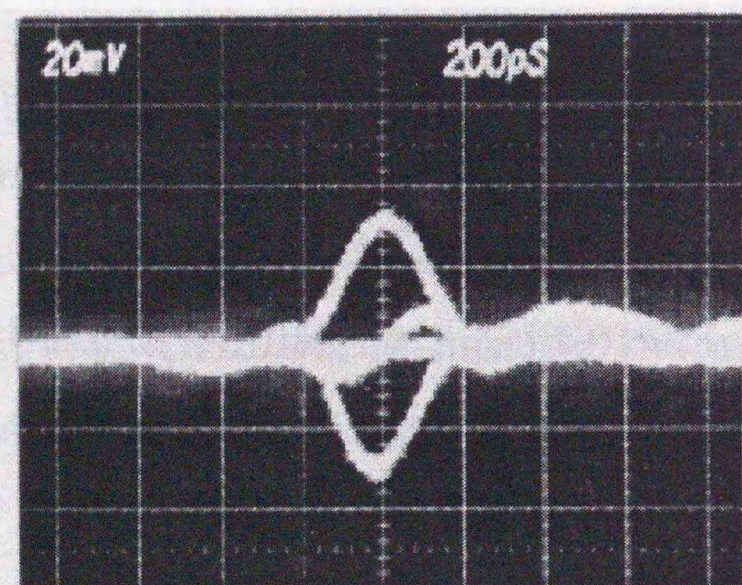
The maximum potential bit rate for the SV signal can be estimated by limiting the circuit imperfection to be negligible ($i_a \rightarrow 0$). Assuming the following parameters: $d = 0.01$, $f_0 = 1.8\text{GHz}$, $N_m = 20$ (dB), $10 \log Q = 23$ (dB), $m = 10$ (10B1C), and $n_f = 10$, the maximum SV signal bit rate is calculated as $f_{sv} = 25$ MHz. Here, the assumed $\text{SNR} = 23$ dB corresponding to the error rate of 10^{-12} .

4.3 Experiment

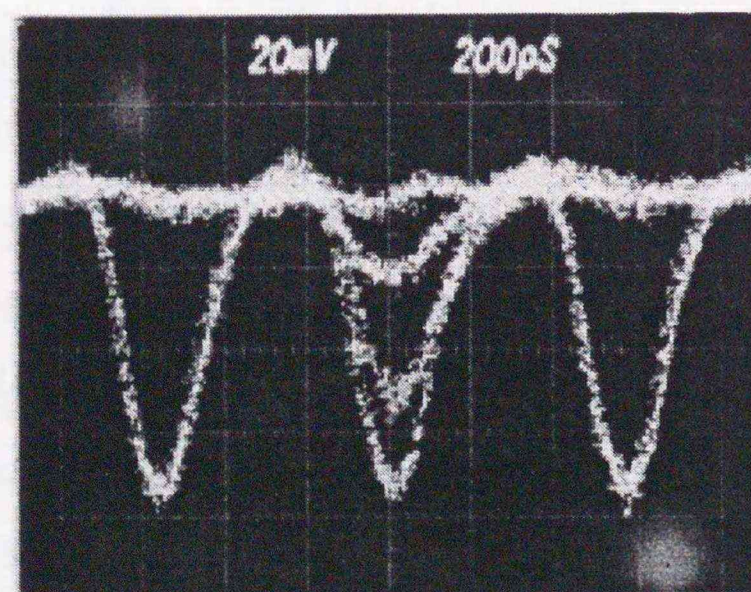
GaAs field-effect transistors (FET's) were used in the modulation and demodulation circuits except for the digital gates and analog gate. The digital gate was composed of OR/NOR small scale integrated circuits (SSI's) fabricated by the self-aligned silicon process technology (SST) [54]. The current switches were also composed of silicon bipolar transistors, but not integrated. Traffic pulse clock frequency was 1.8 GHz, which corresponds to the line clock rate of 1.6 Gbit/s system. Output signal waveforms of the digital gates, analog gate, and current adder are shown in Fig. 4-6(a)-(c), where m is chosen to be 10. A high frequency operational amplifier with 20dB gain and 26 MHz bandwidth (3dB)



(a)



(b)



(c)

Fig. 4-6 Modulating circuit waveform (200 ps/div.) (a) Digital gates output. (b) Modulating analog gate output. (c) Modulated mB1C output sequence.

was used for SV signal amplification. The LPF was a fifth-order Thomson's filter [55].

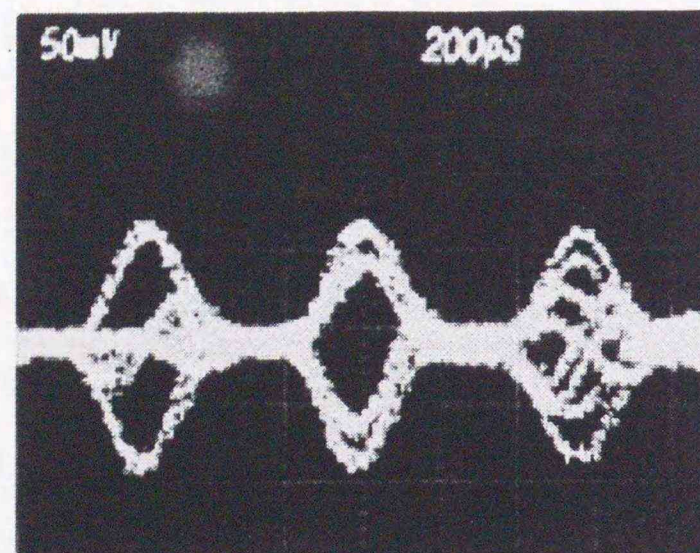
4.3.1 Electrical Performance

A simulated SV signal was 3 Mbit/s pseudorandom sequence of the non-return to zero (NRZ) format. An analog signal could have also been used. In the first step of the SV transmission experiment, the current adder output was fed directly to the differential circuit, the first part of the SV demodulation circuit. Output waveforms from the differential circuit, full-wave rectifier are shown in Fig. 4-7(a) and (b), where the pulse in the middle corresponds to C bit. Demodulated SV signal is shown in Fig. 4-7(c). Error-free operation was obtained beyond 20% modulation depth.

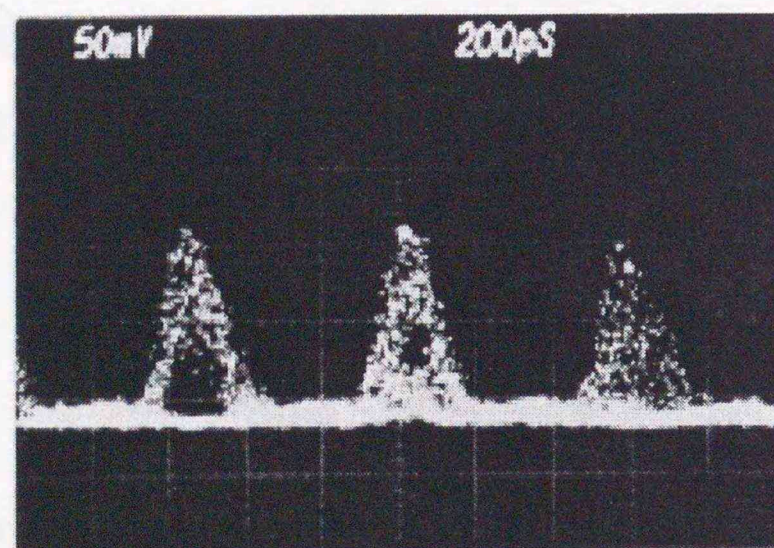
4.3.2 Optical Performance

At transmission experiment through an optical transmitter/receiver was carried out. The transmitter was a Fabry-Perot laser diode of 1.3 μm wavelength. An InGaAs avalanche photodiode (APD) was used for its lower noise as compared to a Ge APD in the gigabit range. Over 40% modulation enabled error-free transmission at an optical power level of -34 dBm. It is shown that in-service SV transmission was feasible because the traffic signal is recoverable at about -30 dBm optical level.

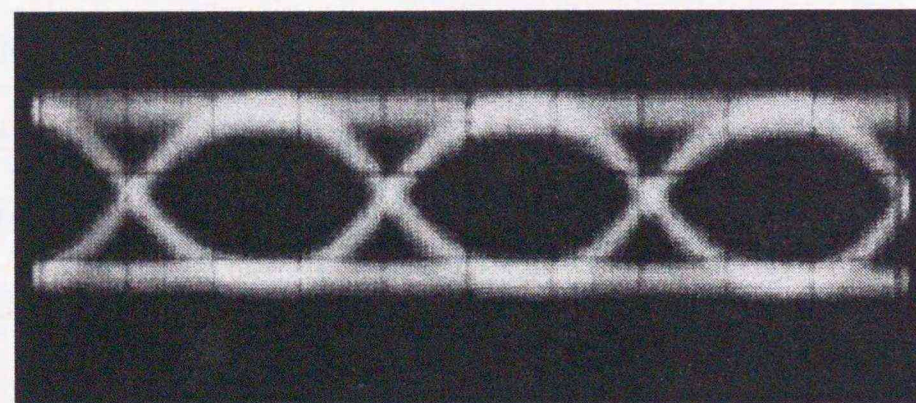
Although this SV signal transmission method includes noise reduction features, some effective noise sources remain in the real circuit. It is important to reduce waveform distortion from circuit nonlinearity and signal leakage at the demodulating analog gate. Since the SV data is applied to the edge of the C bit pulses, small saturation and/or nonlinearity of the analog circuit degrade the



(a)



(b)



(c)

Fig. 4-7 Demodulating circuit waveform. (a) Differential equalizer output (200 ps/div.) (b) Full-wave rectifier output (200 ps/div.) (c) Demodulated SV output (0.1 ms/div.)

SV signal considerably. Imperfection in the demodulating analog gate caused signal leakage from the main data into the SV signal channel. These were the two main effective noise sources in this simulation, and caused the SNR degradation.

4.4 Summary

It has been shown that the C bit modulation for the purpose of SV signal transmission is feasible in gigabit lightwave systems. The C bit, not traffic pulses, is modulated according to SV signal. This technique's advantage is that the main sequence is independent of SV input. This makes it possible to send a relatively high bit rate SV signal ($> \text{Mbit/s}$) within a reasonable circuit scale. The experimental survey proves that error-free transmission can be achieved more than 40% of relative modulation depth, even using a conventional optical repeater with no special care taken to ensure circuit linearity.

Chapter V

Bit-Interleaved Hamming Code for Linearly Repeated Terrestrial Fiber Optic Transmission Systems

5.1 Introduction

One of the important applications of Er-doped fiber amplifiers (EDFAs) is the linear repeaters (L-REPs), because they reduce the number of regenerators in end-to-end transmission and decrease total network cost. Decreasing the number of regenerators yields higher cost effectiveness; however, optical noise accumulation in the L-REPs limits the regenerator span [56]. Overcoming such degradation of signal quality is a crucial issue for implementing longer-span systems. To that end, the forward error-correction (FEC) code is the most promising digital technology for cost-effective and high-quality transmission systems.

Efforts are now being made to apply FEC coding techniques, in optical transmission systems, particularly in submarine transoceanic transmission systems. Usually, multiple-error correcting codes, such as the Bose-Chandhuri-Hocquenghem code [57] or the Reed-Solomon code [58], are selected. Despite

their excellent correcting capability, these codes increase the line rate, which may not be affordable in terrestrial high-speed systems. Line rate increase can be avoided if the check bits can be mapped into the existing unused overhead bytes in the signal frame. Grover and Moore proposed an STS-1 path (52 Mbit/s line rate), equivalently defined VC-3 in SDH, that is encoded in the (6208, 6195) cyclic Hamming code [59]. Thirteen check bits are mapped into the path overhead (POH) auxiliary bytes in the SONET format. It requires no modification to the physical interface or section termination circuits on the line. However, it is not straightforward to apply this proposal to paths other than STS-1. Concatenated Virtual Containers, such as VC-4-Xc ($X=1,4,16$) will be introduced soon to convey ATM cells. Different codes have to be devised for those high-speed paths. Paxal et al. have proposed the Reed-Solomon (524, 522) code [60], where the STM-1 payload is divided into three parts and each fraction is coded in 12 parallel manner. The FEC is independent of path size; however, an FEC circuit must be deployed in each regenerator, and all regenerator circuit would have to be customized. In addition, the code requires different coding circuits for different STM-N ($N>1$) systems, while the decoding process in each regenerator causes the accumulation of significant end-to-end delay. In this chapter, proposed is a single cyclic Hamming code applicable for all SDH signal variants that is compatible with non-FEC systems: it also suppressed delay accumulation [60].

First, we consider where to deploy coding/decoding circuits in SDH network elements: path terminating equipment (PTE) [59], line terminating multiplexer (LTM) or regenerator [60], with the help of Fig. 5-1. The LTM is adopted as the practical solution considering the various SDH levels possible. In the LTM, the FEC circuit can be applied regardless of multiplexing number N because the STM-N frame is a byte-multiplex construction of the STM-1 frame. Moreover, the coding circuit supports all concatenated administrative unit (AU-4-Xc) because they are mapped into STM-1 at the LTM. The codeword is the AU-4

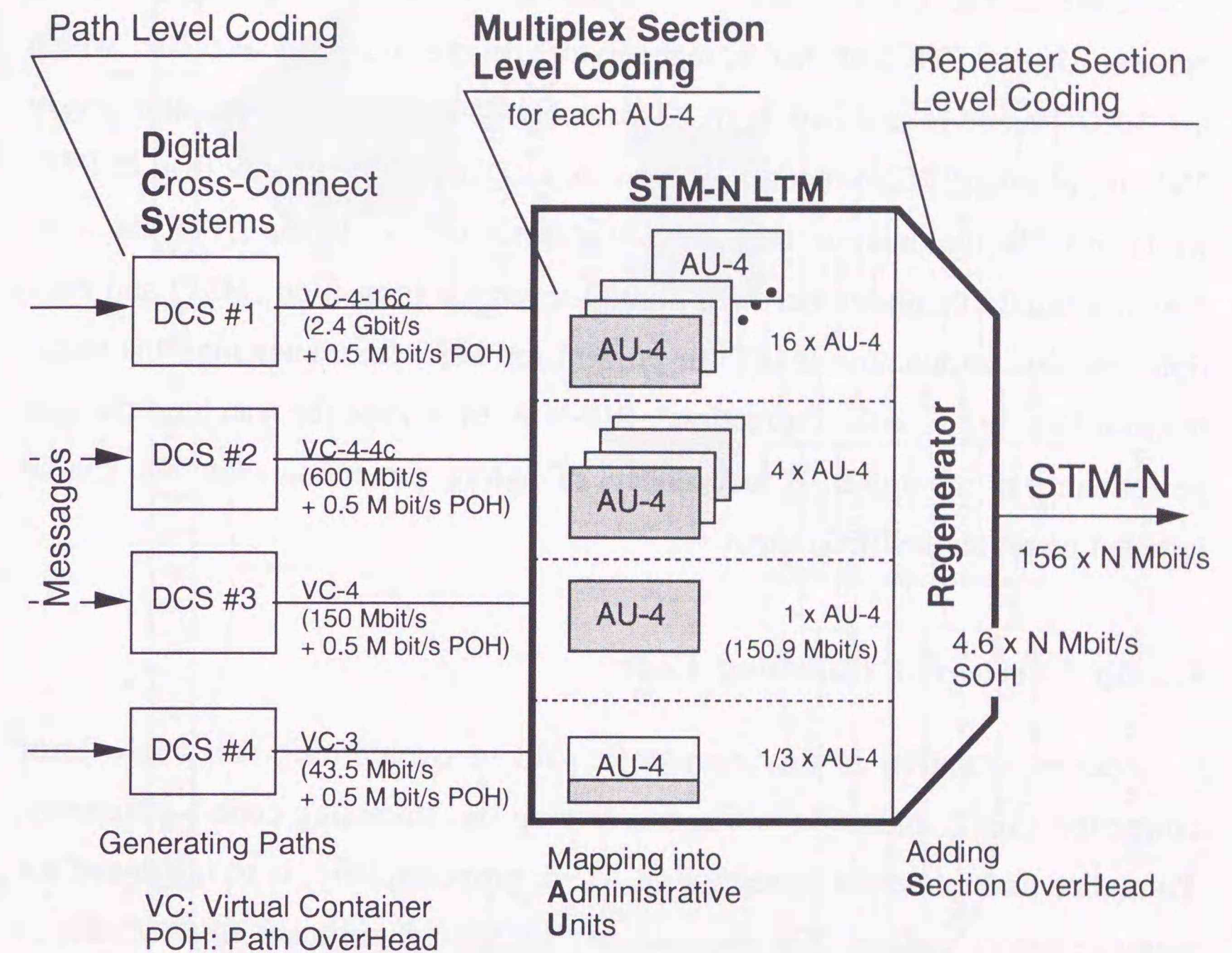


Fig. 5-1 Various possible deployments of transmission coding

itself and the check bits that are mapped into the multiplex section overhead (MSOH). The number of possible check bytes is $24 \times N$ for any STM-N system, so there is no byte shortage for correction purposes as N increases. Placing the FEC code in the LTM offers an intermediate value of delay accumulation between that in PTE and that in regenerator. In the prototypical model which has 5000 km end-to-end path with LTMs every 2500 km and regenerators every 250 km, placing FEC in regenerators yields 110 times larger delay than in PTE, while in LTM the delay is increased by a factor of 10. In the LTM, the FEC function should be placed between multiplex section protection (MSP) and multiplex section termination (MST) to prevent the MSP from switching the transmission line before error correction. BIP- $N \times 24$ is kept for watching the raw performance of the line BER independent of coding; protection switching can be initiated based on the FEC output.

5.2 Bit Interleaved Hamming Code

Special attention in this chapter is paid to overcoming the single error correction (SEC) limitation while maintaining the Hamming code's efficiency. To realize multiple-error correction in AU-4, proposed here, is to interleave the message before coding. For example, all the Most Significant Bits (MSBs) in every target byte constitute one of the eight coding groups, as depicted in Fig. 5-2. An interleaving number 8 was chosen to agree with SDH byte structure. We apply the Hamming code to each group. This allows a single error to be corrected in each group, and up to a single byte error to be corrected overall. While eight coding/decoding pairs are required, the circuits are so simple and small, each circuit operates at just one-eighth the STM-1 bit-rate (19.44 Mbit/s), that all of them can be realized with one CMOS LSI. The byte-based construction of the SDH equipment particularly simplifies signal connection within the LTM.

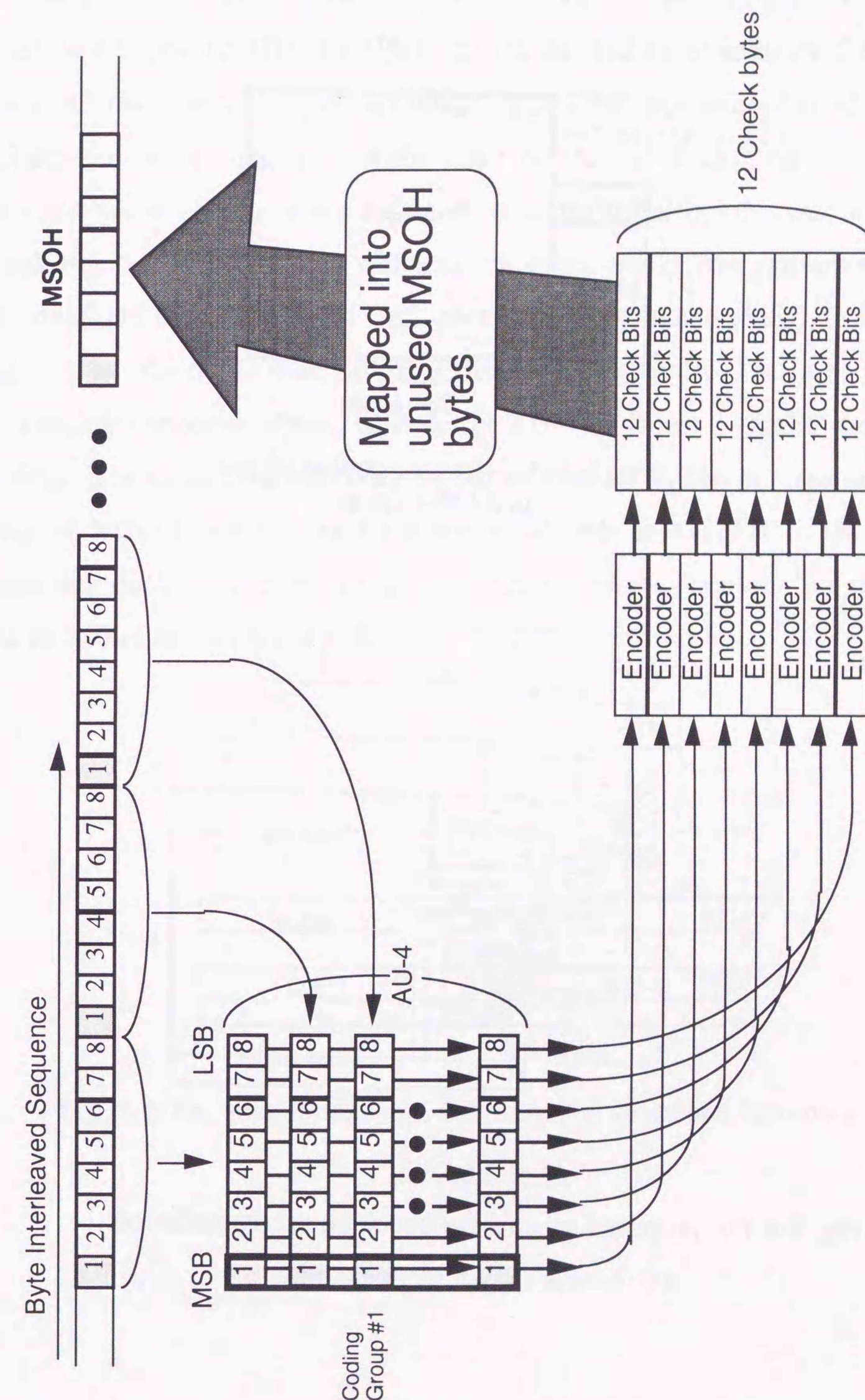


Fig. 5-2 Bit interleaved coding

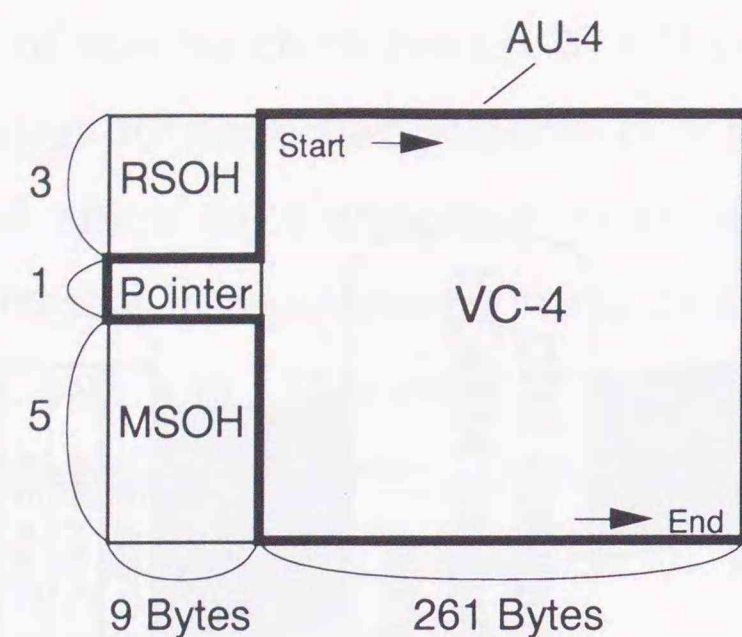


Fig. 5-3 AU-4 as the coding object in STM-1 frame

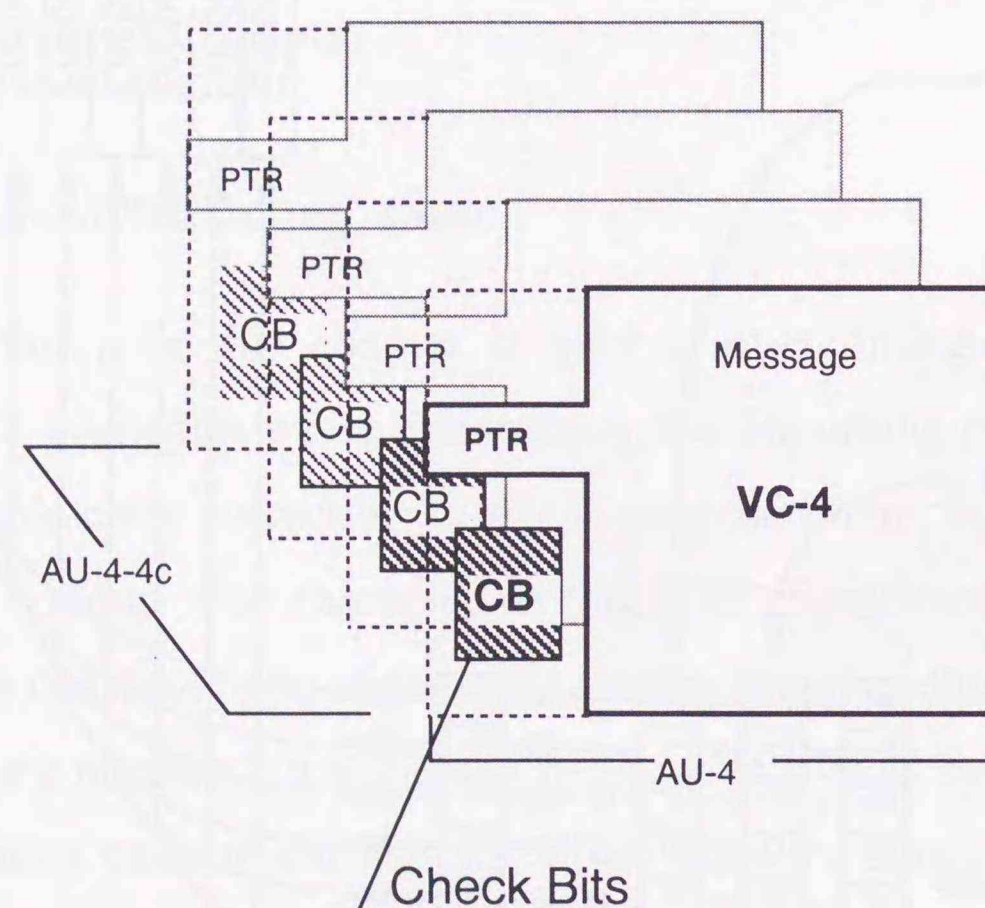


Fig. 5-4 The proposed scheme for check bit accommodation. AU-4-4c is shown as an example

As shown in Fig. 5-3, AU-4 consists of VC-4 payload (261×9 bytes) and 9 bytes of AU-pointer (H1, H2, H3). Since H1, H2 bytes indicate the VC phase, while H3 may carry the payload when the stuffing adjustment is activated, AU-pointer should be included for the error correction block. As a result, each 8 bit-interleaved coding group amounts to 2370 bytes being associated with 12 check bits for each group. Twelve check bytes can be mapped anywhere in the 24 undefined bytes in STM-1 multiplex section overhead (MSOH), as shown in Fig. 5-4, in which AU-4-4c case is depicted as a example. The importance is that the check bit accommodating scheme for AU-4-4c is just a parallel stack of AU-4 coding. The same coder/decoder circuit is used for different sizes or concatenations of VCs. The resulting bit-interleaved code is a $\{2370, 2358\}$ shortened Hamming code, as shown in Fig. 5-5. Circuit block diagrams for the encoding and decoding are shown in Fig. 5-6.

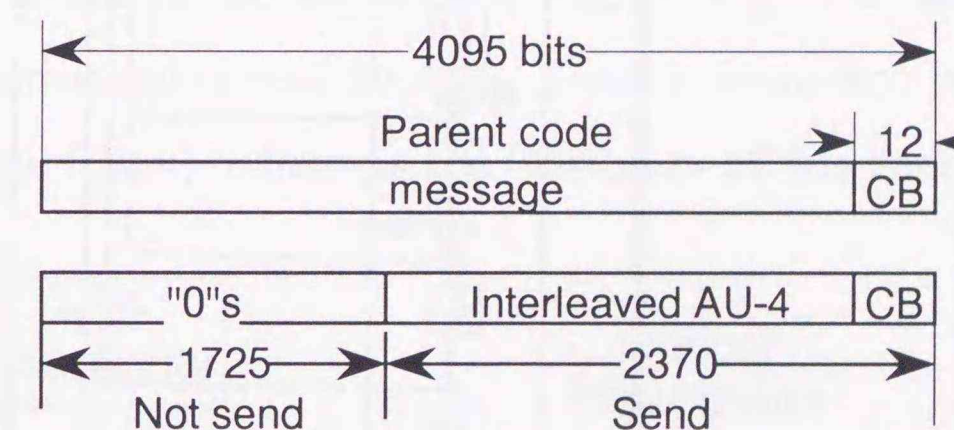
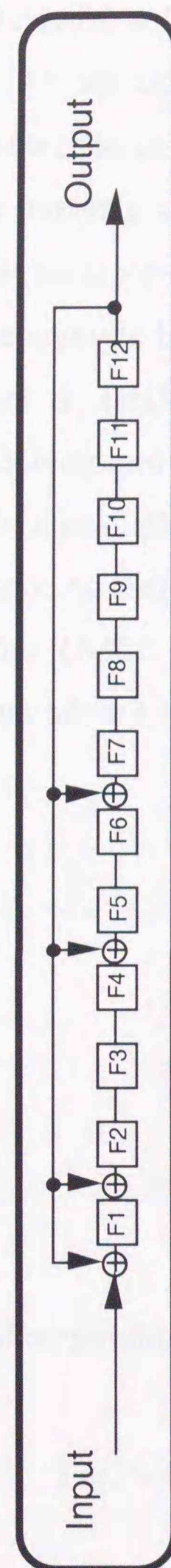


Fig. 5-5 The parent code and the proposed shortened hamming code

ENCODER



DECODER

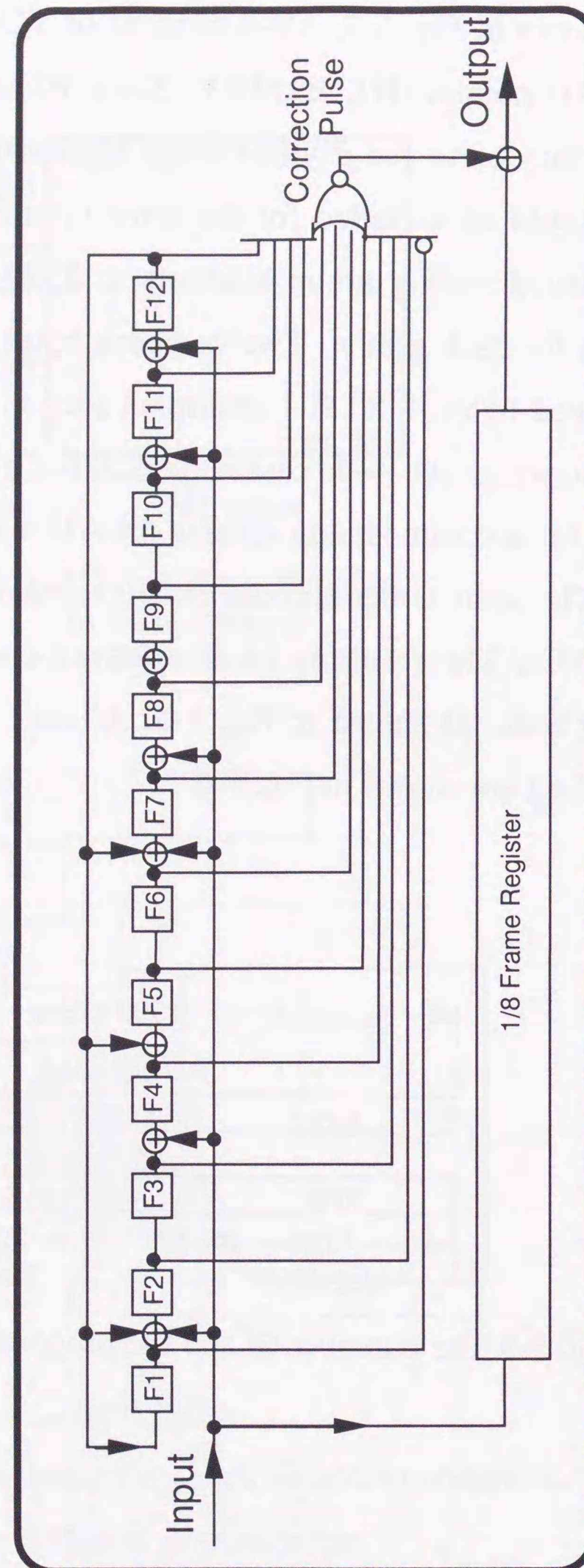


Fig. 5-6 Circuit block diagrams for the encoding (top) and decoding (bottom)

5.3 Experiment

The FEC circuit for a coder and decoder was realized on two connection-changeable integrated circuits or field-programmable gate arrays. The coded AU-4 was merged with the corresponding SOH bytes to form the STM-1 frame at 155.52 Mbit/s. We confirmed FEC performance against amplified spontaneous emission (ASE) noise added to the optical preamplifier in an experimental setup shown in Fig. 5-7, where ASE accumulation through an L-REP chain was simulated. The ASE light with spectral width of 5 nm (FWHM) was generated by an EDFA with no input, and combined with the signal light with a 3dB coupler. At the coupler output, ASE power density was -30 dBm/nm, while the signal power was -28 dBm. Intentional errors were added by reducing the optical input power to the receiver with optical filter of 3 nm FWHM. Figure 5-8 compares the resultant error performances with and without coding. The coding gain of 4.5 dB was obtained at the BER of 10^{-7} . Corrected error rate is plotted as a function of the input error rate in Fig. 5-9, and compared to the theoretically approximated performance. Good coincidence between these two is apparent from the figure, indicating the feasibility of this code.

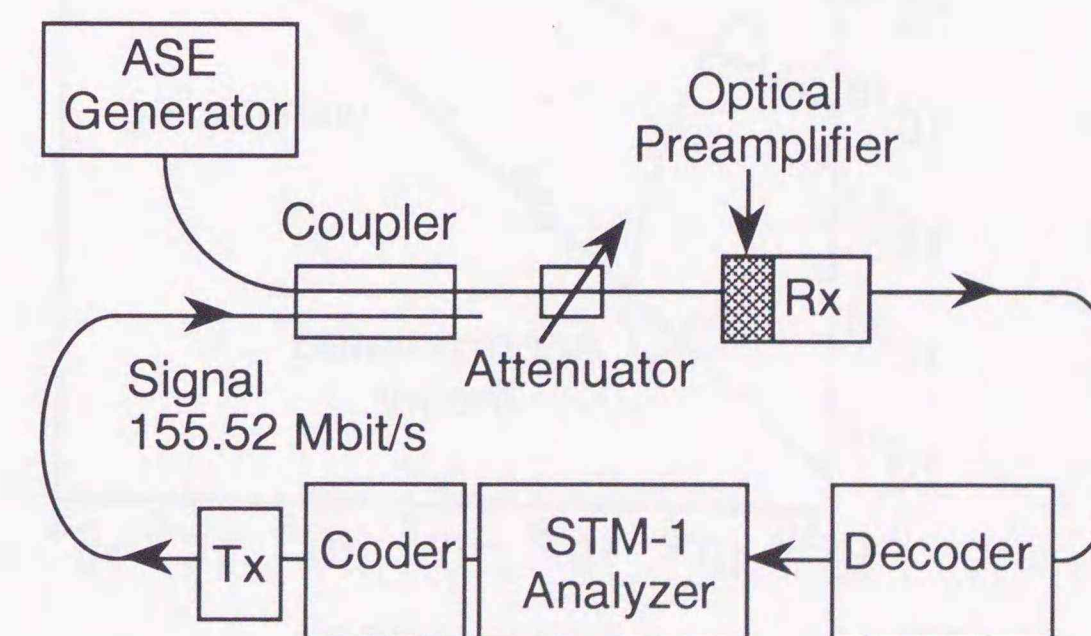


Fig. 5-7 Experimental Setup

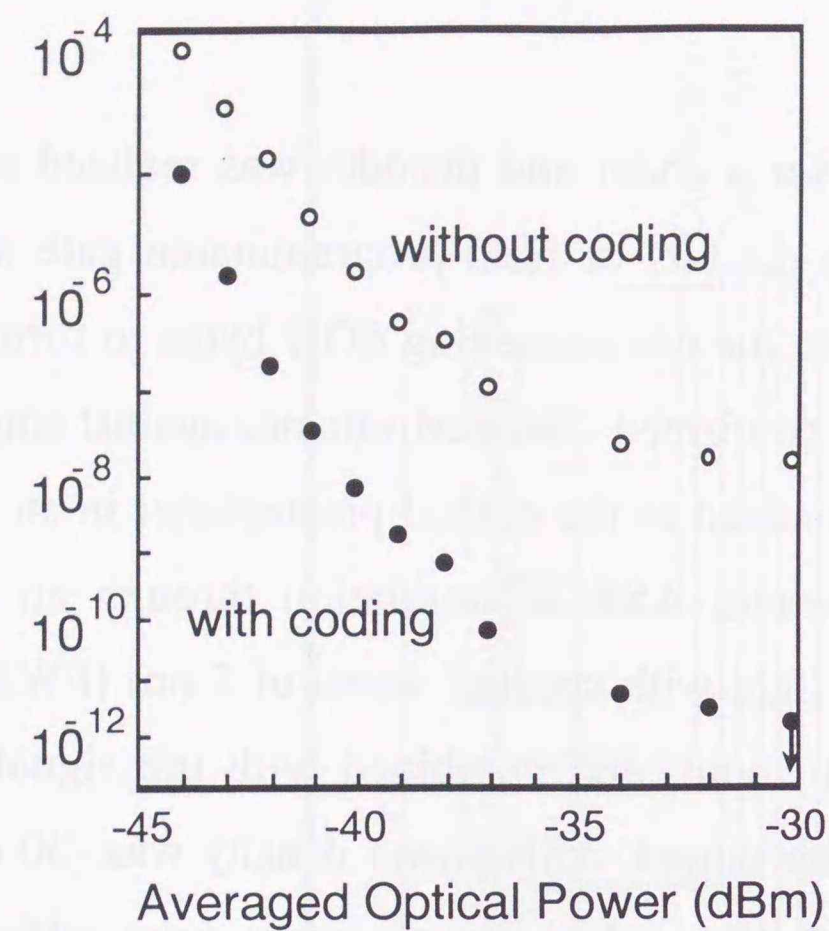


Fig. 5-8 Error performances with and without the coding

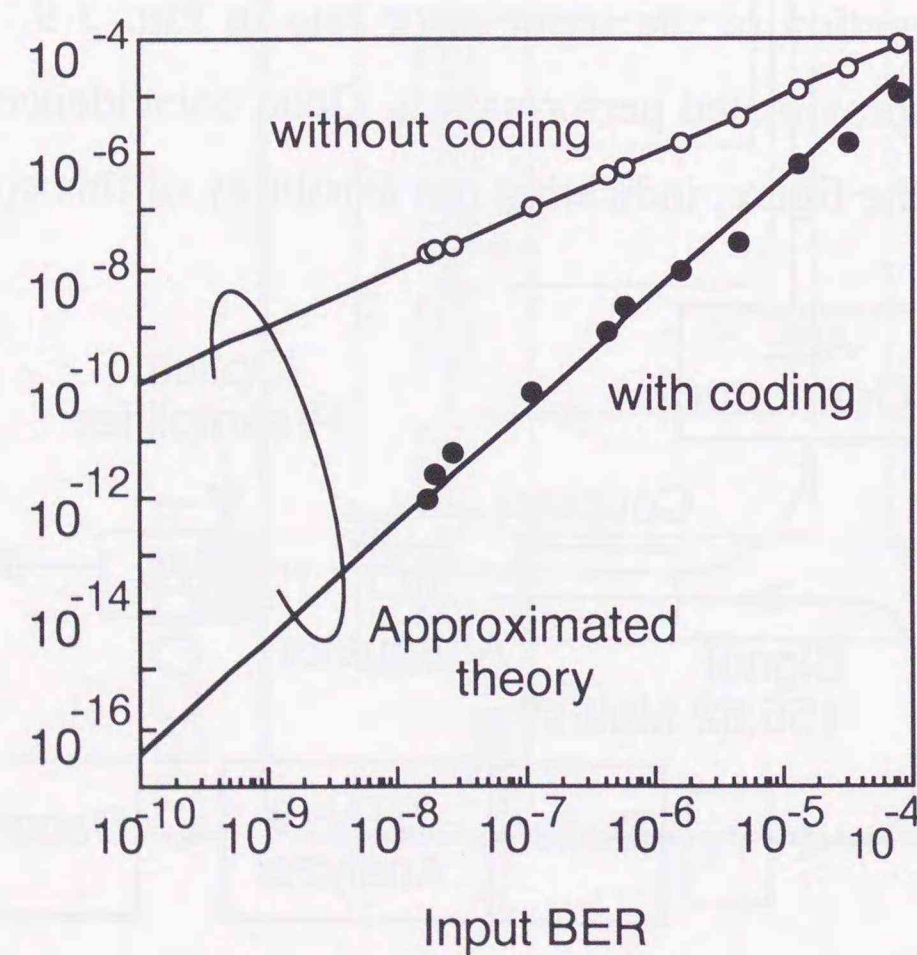


Fig. 5-9 The proposed error correcting code performance

5.4 Summary

An 8 bit interleaved Hamming code for the AU-4 is proposed for SDH line terminating multiplexers. It features up to 1-byte error-correcting capability in the AU-4 by performing (2370, 2358) shortened Hamming code for each interleaved group of the AU-4 bytes. The general code offers the same correction performance regardless of the VC size or STM level by simply arraying the requisite numbers of coders/decoders in parallel. Nothing has to be changed in the configuration of the physical interface circuits for the coded frame. An experimental correction of the ASE error in a 156 Mbit/s optical preamplifier has confirmed FEC performance and feasibility.

Chapter VI

Single-Wavelength Dispersion Measurement for Multiple Fiber Section Connected with Narrow-Band Optical Amplifiers

6.1 Introduction

There have been a number of measurement methods proposed for determining the chromatic dispersion of single-mode fibers. All of them, however, are suitable only for relatively short fibers, or need closed-loop setups. When assessing a long-haul transmission line that includes linear amplifier repeaters, it is important to be able to determine the dispersion value of the entire span with a single measurement. Erbium-doped fiber amplifiers with semiconductor laser pumping are usually employed as the linear repeaters. Although it is possible to measure separately the dispersion value for each repeater span and

sum them up, this technique is too slow and the accumulation of measurement error is significant.

The unidirectional measurement of chromatic dispersion is difficult with conventional techniques, most of which depend on the relative delay between a probe and a reference signal both of which are intensity modulated. There are three difficulties.

The first one is undesirable group delay fluctuation due to random thermal or mechanical disturbances during the measurement. In far-end measurements, the reference signal has to travel the same distance to the receiver as the test signal. To realize accurate measurements, the phase difference between these two signals must be caused only by the chromatic dispersion of the test fiber, not by expansion or compression of the fiber. If an independent reference fiber is used, its effective length must change in the same way as the test fiber under external disturbances. Since this is extremely difficult to achieve, uneven fiber length fluctuation may result in significantly degraded relative-delay measurement accuracy. By using the same fiber and switching transmission from the probe signal to the reference, one can avoid such uneven fluctuation [61]. However, it requires a highly stabilized constant wavelength laser for the reference and needs to be in a closed-loop.

The second difficulty is the spectrum broadening due to frequency chirping resulting from the fiber nonlinearity. In measuring the dispersion of a long fiber, a light source of high output power is needed to overcome fiber attenuation. However, if the source is intensity-modulated, as in almost all of the group-delay measurement techniques, the Self-Phase Modulation (SPM) effect chirps the wavelength at which the dispersion was to have been measured [62]. Nonlinear index of refraction induces optical phase modulation according to the intensity modulation. The SPM affects the intensity waveform in such a manner that the rising portion shifts to the longer wavelength,

while the falling portion shifts the other way. The wavelength is no longer the same over the waveform; therefore, those portions travel with different velocity. Similar line broadening has to be considered if the probe and reference signals are wavelength-multiplexed. Mutual nonlinear phase modulation is called the Cross-Phase Modulation (XPM) and also broadens the optical spectra each other. Consequently, the simple group-delay measurement is spoiled.

The third problem is the restricted bandwidth of the transmission line. Optical filters of nanometer bandwidth are usually placed in each amplifier repeater in order to block its undesirable spontaneous emission. Conventional measurement methods using multiple-wavelength light sources are unsuitable because only a part of the probe spectra will pass through the filters. A devised version intended for narrow (4 nm) window requires the 36 data points to suppress the fitting error [61]. The wavelength step has to be very fine; therefore, both measurements for the group delay and wavelength have to be accurate and simultaneous.

As mentioned above, conventional methods for measuring chromatic dispersion are not suitable for long spans, especially if linear amplifier repeaters are used. To enable the precise and simple measurement of dispersion, a new method must be created that does not use a reference signal. The probe light has constant intensity, at least in the beginning of the propagation, where the optical power may be strong enough to cause the SPM effects. The bandwidth of the probe light must match that of the optical filter(s) in the fiber system being measured.

We show in this chapter that the PM-AM conversion effect, using a sinusoidally phase modulated but with constant intensity light from a single-mode laser, can be successfully used for determining the chromatic dispersion of a concatenated fiber section.

6.2 Theory

6.2.1 PM-AM Conversion

The principle itself is well-known. The amplitude at the phase modulation frequency normalized by the constant component is determined in [63] and [64]. To determine the measurable limit, we extend their result by considering the source coherence, or spectral linewidth, given as follows.

Consider a single-mode fiber with an optical source whose electric field at the input is expressed as

$$e_{in}(t) = E_0 \exp[j(\omega_0 t + b_m \sin \omega_m t + \phi(t))], \quad (6-1)$$

where E_0 is the amplitude, and ω_0 is the frequency, $e_{in}(t)$ is phase modulated by $b_m \sin \omega_m t$, where b_m is the modulation index and ω_m is the modulation frequency. The input field can be expressed in terms of a Fourier spectrum,

$$E_{in}(\omega) = \frac{1}{2\pi} E_0 \sum_n J_n(b_m) \int \exp\{j\phi(t)\} \exp\{-j(\omega - (\omega_0 + n\omega_m))t\} dt \quad (6-2)$$

The phase noise $\phi(t)$ is assumed to have zero-mean, and be white; i. e.,

$$\langle \phi(t) \rangle = 0, \quad (6-3)$$

$$\langle \dot{\phi}(t) \dot{\phi}(t + \tau) \rangle = 2\delta\omega \delta(\tau), \quad (6-4)$$

where $\delta\omega$ represents the spectral half-width at half-maximum of e_{in} , “.” denotes the time derivative, and $\langle \dots \rangle$ denotes the ensemble average. The output field after traveling a dispersive fiber of length L can be written as a function of time in a convolution form:

$$e_{out}(\omega) = E_0 \sum_n J_n(b_m) \exp\left\{j\left(\beta + \beta'(\omega - \omega_0) + \beta''(\omega - \omega_0)^2\right)L\right\} \cdot \hat{\Phi}(\omega - (\omega_0 + n\omega_m)), \quad (6-5)$$

where β is the propagation constant of the fiber, and each prime “’” is for $\frac{\partial}{\partial \omega}$, and

$$\hat{\Phi}(\omega - (\omega_0 + n\omega_m)) = \frac{1}{2\pi} \int \exp\{j\phi(t)\} \exp\{-j(\omega - (\omega_0 + n\omega_m))t\} dt. \quad (6-6)$$

The second derivative of the propagation constant β is inversely proportional to the "chromatic dispersion," which is defined the group delay as a function of wavelength. In order to obtain output field expression or Fourier transform of (6-5), we use the convolution theorem

$$f(t) \otimes g(t) = F^{-1}[F[f(t)] \cdot F[g(t)]], \quad (6-7)$$

where $\otimes$ denotes convolution integral, $F[\cdot]$ represents Fourier transform. We also evoke

$$\begin{aligned} f(t) &= \int \exp\left\{j\left(\beta + \beta'(\omega - \omega_0) + \beta''(\omega - \omega_0)^2\right)L\right\} e^{j\omega t} d\omega \\ &= e^{j(\beta L + \omega_0 t)} \int \exp\left\{j\frac{1}{2}\beta''Lv^2 + (t + \beta'L)v\right\} dv \\ &= \sqrt{\frac{2\pi}{\beta''L}} e^{j(\beta L + \omega_0 t)} \exp\left\{j\frac{(t + \beta'L)^2}{2\beta''L}\right\}, \end{aligned} \quad (6-8)$$

where $v = \omega - \omega_0$, and assumed $\beta'' \neq 0$ and

$$\begin{aligned} g(t) &= F^{-1}[\hat{\Phi}(\omega - (\omega_0 + n\omega_m))] \\ &= \exp\{j((\omega_0 + n\omega_m)t + \phi(t))\}. \end{aligned} \quad (6-9)$$

The output field as a function of time in a convolution form

$$e_{out}(t) = E_0 \sqrt{\frac{2\pi}{\beta''L}} \sum_n J_n(b_m) \int \exp \left[j \left\{ \beta'L + \omega_0 \tau + \frac{(t + \beta'L)^2}{2\beta''L} \right\} \right] \cdot \exp \left[j \left\{ (\omega_0 + n\omega_m)(t - \tau) + \phi(t - \tau) \right\} \right] d\tau. \quad (6-10)$$

The receiver transforms it into a photocurrent through square-law detection, resulting in

$$|e_{out}(t)|^2 = E_0^2 \sum_n J_n(b_m) \sum_k J_k(b_m) \left(\frac{2\pi}{\beta''L} \right) \exp \{ j(n-k)\omega_m t \} \cdot \int d\tau_1 \int d\tau_2 \exp \left\{ j \left(\frac{(\tau_1 + \tau_2)(\tau_2 - \tau_1)}{2\beta''L} + \frac{\beta'}{\beta''}(\tau_2 - \tau_1) \right) \right\} \cdot \exp \{ j(-k\omega_m \tau_2 + n\omega_m \tau_1 + \phi(t - \tau_2) - \phi(t - \tau_1)) \}, \quad (6-11)$$

The double integral can be simplified if we take new coordinate

$$\begin{aligned} \tau_1 + \tau_2 &= \xi \\ \tau_1 - \tau_2 &= \eta. \end{aligned} \quad (6-12)$$

The double integral becomes

$$\begin{aligned} & \int d\xi \int d\eta \exp \left\{ j \left(\frac{\eta}{2\beta''L} - \frac{(n-k)}{2} \omega_m \right) \xi \right\} \\ & \cdot \exp \left\{ j \left(\frac{\beta'}{\beta''} \eta + \frac{(n+k)}{2} \omega_m \eta + \phi(t - \frac{\eta}{2}) - \phi(t + \frac{\eta}{2}) \right) \right\} \\ & = \frac{\beta''L}{\pi} \left[\exp \left\{ j \left(\beta'L(n-k)\omega_m + \frac{1}{2}\beta''L(n^2 - k^2)\omega_m^2 \right) \right\} \right. \\ & \quad \left. + \phi(t - \frac{1}{2}\beta''L(n-k)\omega_m) - \phi(t + \frac{1}{2}\beta''L(n-k)\omega_m) \right]. \end{aligned} \quad (6-13)$$

The first term in the exponent simply means a phase evolution of the converted intensity corrugation, therefore it is neglected hereafter. The phase noise portion can be rewritten as

$$\exp \left[j \left\{ \phi \left(t - \frac{1}{2} \tau_{n-k} \right) - \phi \left(t + \frac{1}{2} \tau_{n-k} \right) \right\} \right], \quad (6-14)$$

where, $\tau_{n-k} = \beta''L(n-k)\omega_m$, denoting the delay between beating spectra. Our measurement technique is based on beat generation between the adjacent Bessel components which originate from the phase modulation, *i. e.* $n-k = \pm 1$. The first-order sideband signal can be expressed as

$$I_1(t) = 4E_0^2 \sum_{i \geq 0} J_i(b_m) J_{i+1}(b_m) \sin \left\{ \left(\frac{1}{2} \right) (2i+1) \beta''L \omega_m^2 \right\} \exp \left[-j \int_{t-\frac{1}{2}\tau_1}^{t+\frac{1}{2}\tau_1} \dot{\phi}(\theta) d\theta \right]. \quad (6-15)$$

To obtain the power spectrum $\hat{S}(\omega)$ of the first sideband of the beating photocurrent, we calculated its Fourier pair, or autocorrelation function $R(u)$ as

$$\begin{aligned} R(u) &= \langle I(t) I^*(t+u) \rangle \\ &= \left[4E_0^2 \sum_{i=0} J_i(b_m) J_{i+1}(b_m) \sin \left\{ \left(\frac{1}{2} \right) (2i+1) \beta''L \omega_m^2 \right\} \right]^2 \\ & \times \left\langle \exp \left[j \left\{ \int_{t+u-\frac{1}{2}\tau_1}^{t+u+\frac{1}{2}\tau_1} \dot{\phi}(\theta) d\theta - \int_{t-\frac{1}{2}\tau_1}^{t+\frac{1}{2}\tau_1} \dot{\phi}(\theta) d\theta \right\} \right] \right\rangle \end{aligned} \quad (6-16)$$

Under the ergodic assumption, here we take the ensemble average instead of the infinite time average. The last term, which reflects the coherency of the light source, can be approximated as (see Appendix)

$$\exp[-2\delta\omega |u|] \quad \text{for } |u| \leq \tau_1 \quad (6-17)$$

$$\exp[-2\delta\omega \tau_1] \quad \text{for } |u| > \tau_1. \quad (6-18)$$

The same calculation has been described in reference [65] with relatively rough explanation. As a consequence, we obtain the power spectrum

$$\hat{S}(\omega) = A_1^2 A_2, \quad (6-19)$$

where

$$A_1 = 4E_0^2 \sum_{i=0} J_i(b_m) J_{i+1}(b_m) \sin\left\{\frac{1}{2}(2i+1)\beta''L\omega_m^2\right\}$$

$$A_2 = e^{-2\delta\omega \tau_1} \cdot \left[\delta(\omega - \omega_m) - \frac{\sin[\tau_1(\omega - \omega_m)]}{\pi(\omega - \omega_m)} \right]$$

$$+ \frac{1}{\pi} \frac{2\delta\omega}{(\omega - \omega_m)^2 + (2\delta\omega)^2}$$

$$\cdot \left[1 - e^{-2\delta\omega \tau_1} \left\{ \cos[\tau_1(\omega - \omega_m)] - \frac{(\omega - \omega_m)}{2\delta\omega} \sin[\tau_1(\omega - \omega_m)] \right\} \right].$$

The term A_2 of (6-19) is constituted of the sharp beat spectrum depressed by the factor of $\exp(-2\delta\omega \tau_1)$ and the Lorentzian noise spectrum. Obviously, when the linewidth is extremely narrow, the result becomes that obtained by Chraplyvy *et al.* [63].

Normalized amplitude $\left(\lim_{\omega \rightarrow \omega_m} |\hat{S}(\omega)|^{1/2} / E_0^2 \right)$ is plotted in Fig. 6-1 for four linewidth values $(2\delta\omega / 2\pi)$, from 10 MHz to 1 GHz, showing the influence of source coherence. The horizontal axis is total dispersion, which is the product of fiber length (km) and the conventional dispersion value D (ps/(km nm)) defined as

$$D = -\frac{2\pi c}{\lambda^2} \beta'', \quad (6-20)$$

where c is the velocity of light, and λ is its wavelength in a vacuum. The perfect-coherent case is also indicated with a dashed line. Since the normalized amplitude is a sine function of $\beta''L\omega_m^2$, the observed value of the amplitude indicates multiple values of total dispersion. It is better to limit our attention to the first quadrant, namely,

$$\beta''L\omega_m^2 < \pi, \quad (6-21)$$

to simplify the measurement. This boundary is plotted with a dashed line in Fig. 6-2, when the modulation frequency is 3 GHz as in the experiment described later. If the total dispersion of the fiber under test is completely unknown, the initial measurement should be with a low modulation frequency, which gives negligible beat amplitude, and then raise the modulation frequency up to the value at which sufficient beat amplitude can be obtained. Inequality (6-21) sets no practical limitation on the measurable total dispersion $\beta''L$ if the modulation frequency is tunable. On the other hand, the coherence or finite linewidth of the light source reduces the normalized beat amplitude. As is shown in Fig. 6-1, error can be avoided by the use of laser diodes with 10 MHz linewidth or less. Such devices are commercially available at 1.5 μm wavelength window, which is used in long-distance communication systems. The use of such a laser is preferred to referring to a complex theoretical curve which takes into account the coherence factor B in (6-19), in converting the observed amplitude to the total dispersion value. Three parameters; fiber length, dispersion value, and source linewidth, have a trade-off relation, as shown in Fig. 6-2, where the 10% reduction contour is plotted for various linewidths. It shows that if we try to measure a 1000 km dispersion-unshifted fiber at 1.55 μm ($D \cong 15$ ps/(km nm)), the linewidth must be narrower than 100 MHz. The modulation frequency of 3 GHz at 1.55 μm , yields a first maximum of the sine function at about $DL = 7000$ ps/nm, which means that

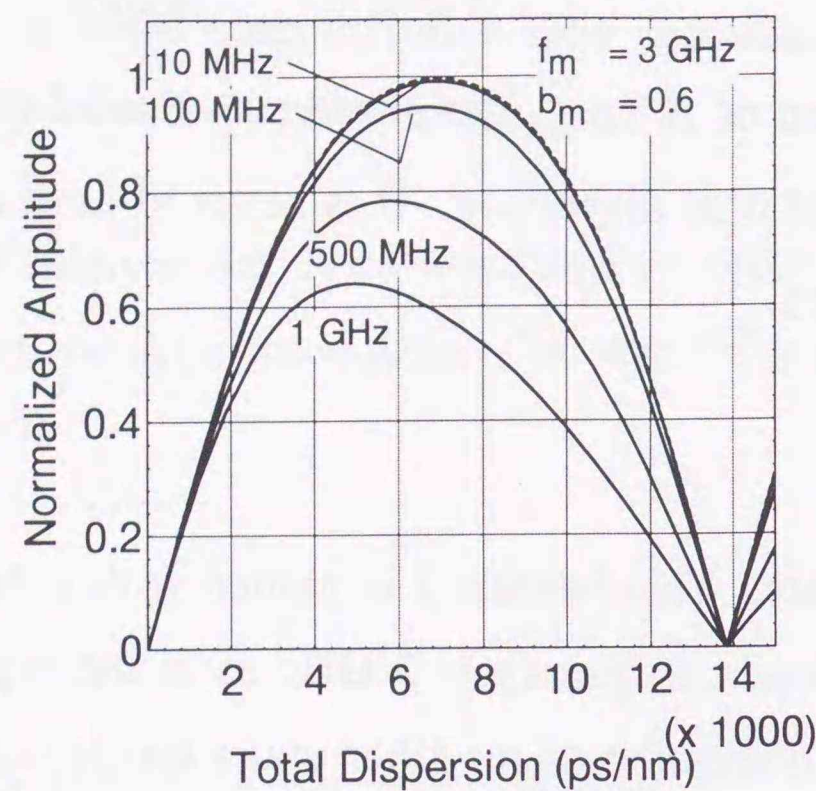


Fig. 6-1 Normalized beat amplitude as a function of total dispersion. Different linewidths from 10 MHz to 1 GHz are assumed for the light source. Dashed line indicates the case of perfect coherence ($2\delta\omega/2\pi = 0$).

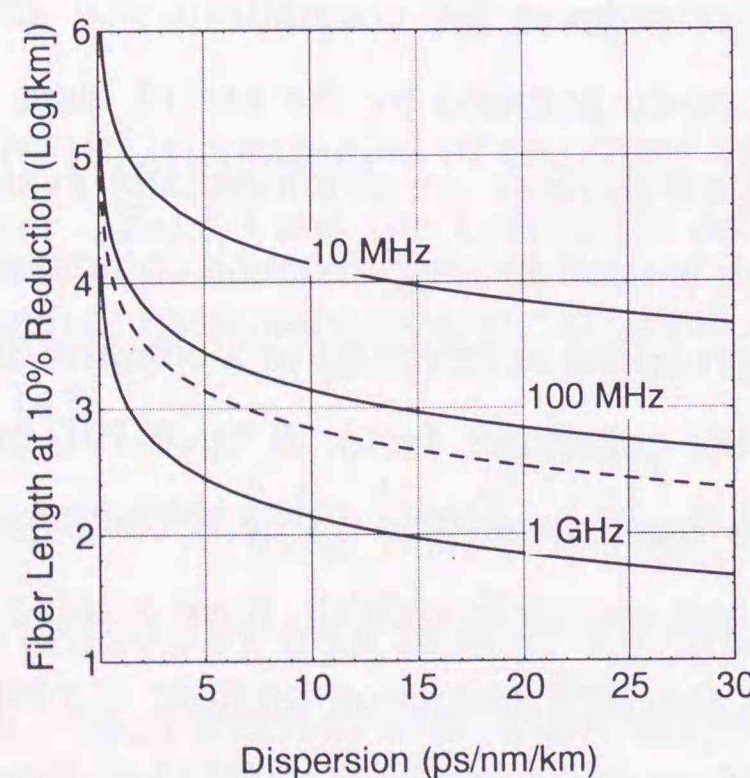


Fig. 6-2 Fiber length at which the normalized amplitude is reduced by 10% due to finite linewidth of the light source. Dashed line indicates the maximum of the normalized amplitude for the perfect coherent case with the modulation frequency of 3 GHz.

3500 km dispersion-shifted fibers ($D = 2$ ps/(km nm)) can be measured. Since larger dispersion values or longer lengths permit lower modulation frequency, the proposed method is applicable to any type of installed fiber systems even if they contain optical amplifiers.

6.2.2 Signal-to-Noise Ratio (SNR)

In the proposed measurement scheme, the signal amplitude modulation can be thought to be a coherent beat between the sidebands due to the phase modulation. The amplitude modulation power S grows as the total dispersion (DL [ps/nm]) increases, according to A in (6-19).

$$S = \left[4G \langle N_{in} \rangle J_0(b_m) J_1(b_m) \sin \left\{ \frac{\beta'' L \omega_m^2}{2} \right\} \right]^2 \exp[-2\delta\omega\tau_1], \quad (6-22)$$

where only the first term ($i = 0$) is considered for approximation.

On the other hand, "Noise" N includes shot noise by both (i) laser light (time-averaged) and (ii) amplified spontaneous emission (ASE) from the amplifiers, (iii) Signal-ASE beat noise, (iv) ASE-ASE beat noise, and (v) *incoherent* beat between the sidebands. Along with the signal increase, an effective noise component due to incoherent beating also increases. It is given in the second term of A_2 in (6-19) by the Lorentzian spectrum being centered at the modulation frequency. Since it has only twice the linewidth of the light source, the incoherent beat note plays as a "narrow-band noise," reducing the SNR as the dispersion or source linewidth becomes large. Receiver's thermal noise can be neglected if the gain G is large [66]. Shot noise power is proportional to the mean value of the photon number per second incident on the photodetector,

$$\langle N_{out} \rangle = \langle N_{in} \rangle + (k+1)(G-1)\alpha n_{sp} m_i \Delta f, \quad (6-23)$$

where the first term is the input laser photon number, which is assumed to be reproduced by the k repeaters and an optical preamplifier located in front of the detector. The second term represents the ASE accumulation, where fiber loss $\alpha = 1/G$, the population inversion parameter n_{sp} , effective number of transverse modes or polarization modes m_t , and bandwidth Δf of the optical bandpass filter are assumed to be equal for all amplifiers for simplicity. Using (6-23), noise components can be expressed in the order of appearance above as

$$\begin{aligned} N_{shot}^{sig} &= 2B \langle N_{in} \rangle \\ N_{shot}^{ASE} &= 2B(k+1)(G-1)\alpha n_{sp} m_t \Delta f \\ N_{beat}^{sig-ASE} &= 2B \cdot 2(k+1)(G-1)\alpha^2 n_{sp} \langle N_{in} \rangle \\ N_{beat}^{ASE-ASE} &= 2B \cdot 2(k+1)^2 (G-1)^2 \alpha^2 n_{sp}^2 m_t \Delta f \\ N_{incoh} &= \left[4 \langle N_{in} \rangle J_0(b_m) J_1(b_m) \sin \left\{ \frac{\beta'' L \omega_m^2}{2} \right\} \right]^2 (1 - e^{-2\delta\omega\tau_1}) b, \end{aligned} \quad (6-24)$$

where B is electrical bandwidth of the receiver circuit, and b is the effective noise bandwidth

$$b = \int_{-\infty}^{\infty} \frac{1}{\pi} \frac{2\delta\omega}{(\omega - \omega_m)^2 + (2\delta\omega)^2} d\omega. \quad (6-25)$$

Figure 6-3 shows normalized power traces of those broadband noises and the total noise as functions of the number of linear repeaters. The parameters assumed are listed in Table 6-1. In that particular case, signal shot noise surpasses other noises for up to 10 repeaters, and ASE-ASE beat noise dominates at 100 repeaters and beyond. Figure 6-4 shows the overall SNR as a function of the total dispersion, where the number of the repeaters is assumed to be 10, and we again use the parameters in Table 6-1. The rising slope can be improved by either narrowing the receiver's electrical bandwidth or

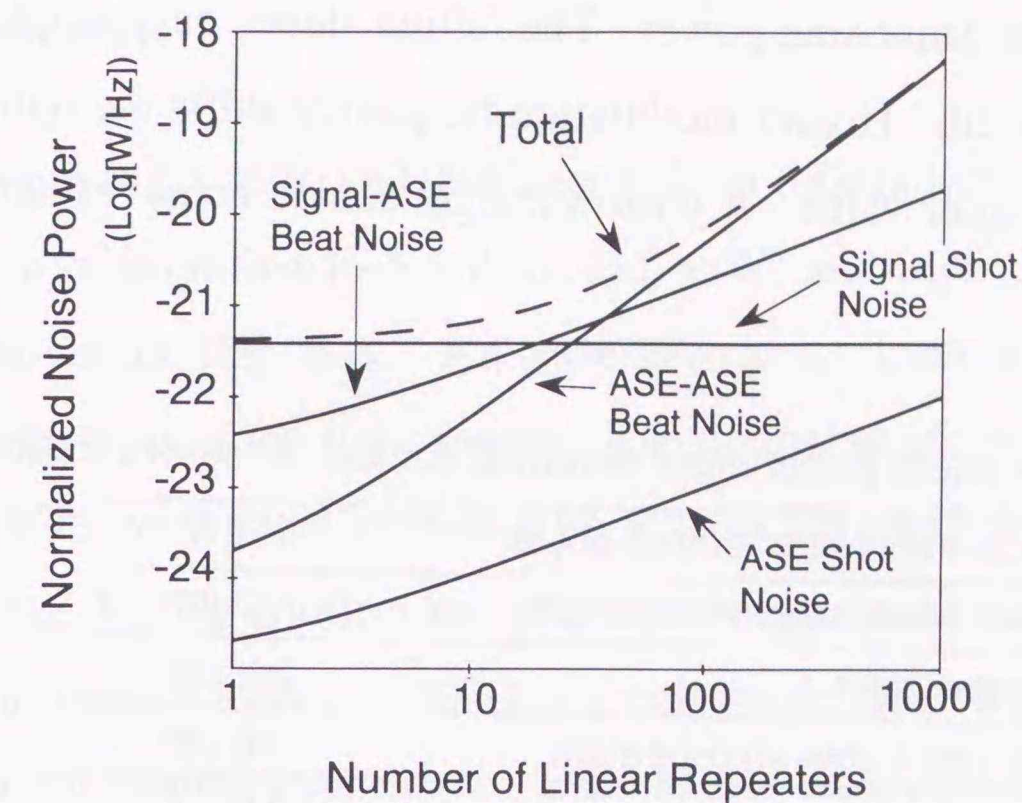


Fig. 6-3 Noise power normalized with frequency (W/Hz) against number of optical amplifier repeaters.

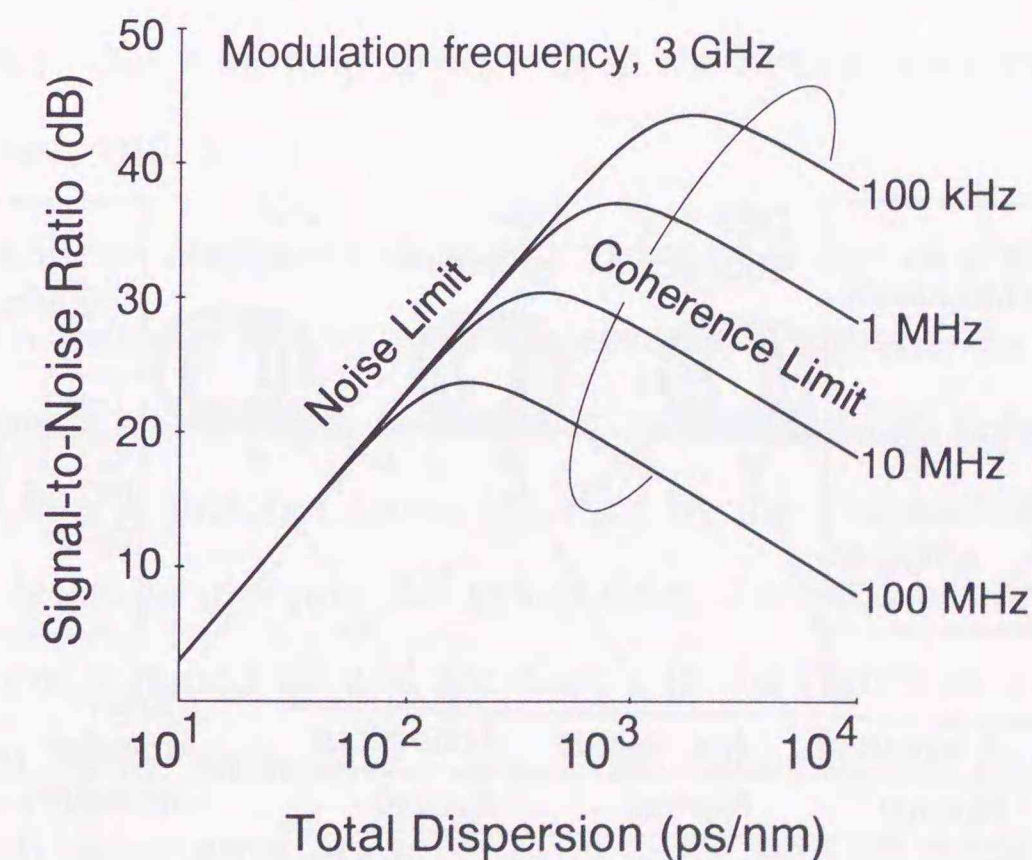


Fig. 6-4 Overall SNR as a function of the total dispersion for 10-repeater system with the parameters in Table 6-1.

increasing the optical launching power. The falling slopes depend on the optical source linewidth. Higher modulation frequency shifts the optimum dispersion to the smaller value. A narrower light source achieves better SNR.

Table 6-1 System parameters assumed for the noise evaluation

Optical transmitter output power	1 mW
Optical transmitter wavelength	1.5 μm
Repeater gain	20. dB
Fiber loss between repeaters	-20 dB
Population inversion parameter	2.5
Number of the lateral mode	2
Optical bandpass-filter width	3 nm
Modulation frequency	3 GHz
Receiver bandwidth	15 GHz

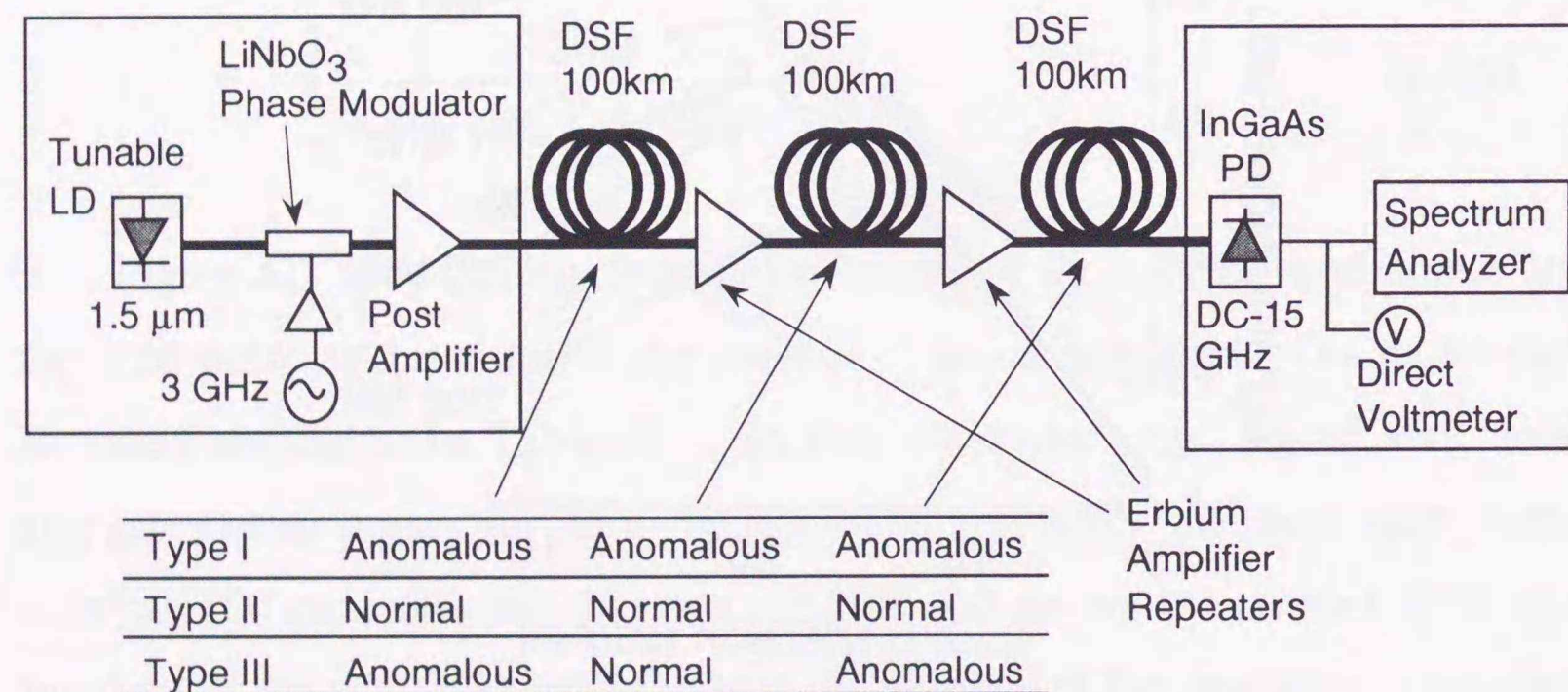


Fig. 6-5 Experimental setup

6.3 Experiment

Single-mode phase-modulated laser light at 3 GHz with the modulation index of 0.6 was amplified to 5.5 dBm (3.5 mW) and injected into a test fiber line, as shown in Fig. 6-5. An external-cavity laser diode of 100 kHz linewidth was used as the light source. The optical phase modulator used here was a LiNbO₃ waveguide crystal with 4.7 dB loss and the half-wavelength voltage of 9.8 V. To confirm the proposed method's suitability for concatenated fibers, three bobbins of 100 km single-mode fiber were linked together via two erbium amplifier repeaters. The input to each amplifier was about -16 dBm, and their output was about +5 dBm. Three dispersion cases were tested, namely all anomalous dispersion fiber (Type I), all normal dispersion fiber (Type II), and a combination of anomalous and normal dispersion fibers (Type III), for a total length of 300 km each. The output light was detected with a broadband (DC - 15 GHz) photodiode followed by a spectrum analyzer and a direct voltmeter. No nonlinear distortion in the optical spectra was observed on an optical spectrum analyzer.

The normalized amplitude depicted in Fig. 6-6 gives absolute values of dispersion. To distinguish the polarity, normal or anomalous dispersion, we scanned the source wavelength by 3 nm. As shown in Fig. 6-6, the dispersion data obtained clearly matches those obtained by the conventional method that uses multiple lasers to measure the group delay for each bobbin. The dispersion values were summed up and are shown in the figure as the shaded belts indicating ± 90 ps/nm ambiguity. Since the results of the proposed method have at most 50 ps/nm error, its superiority to the sum-up method is obvious.

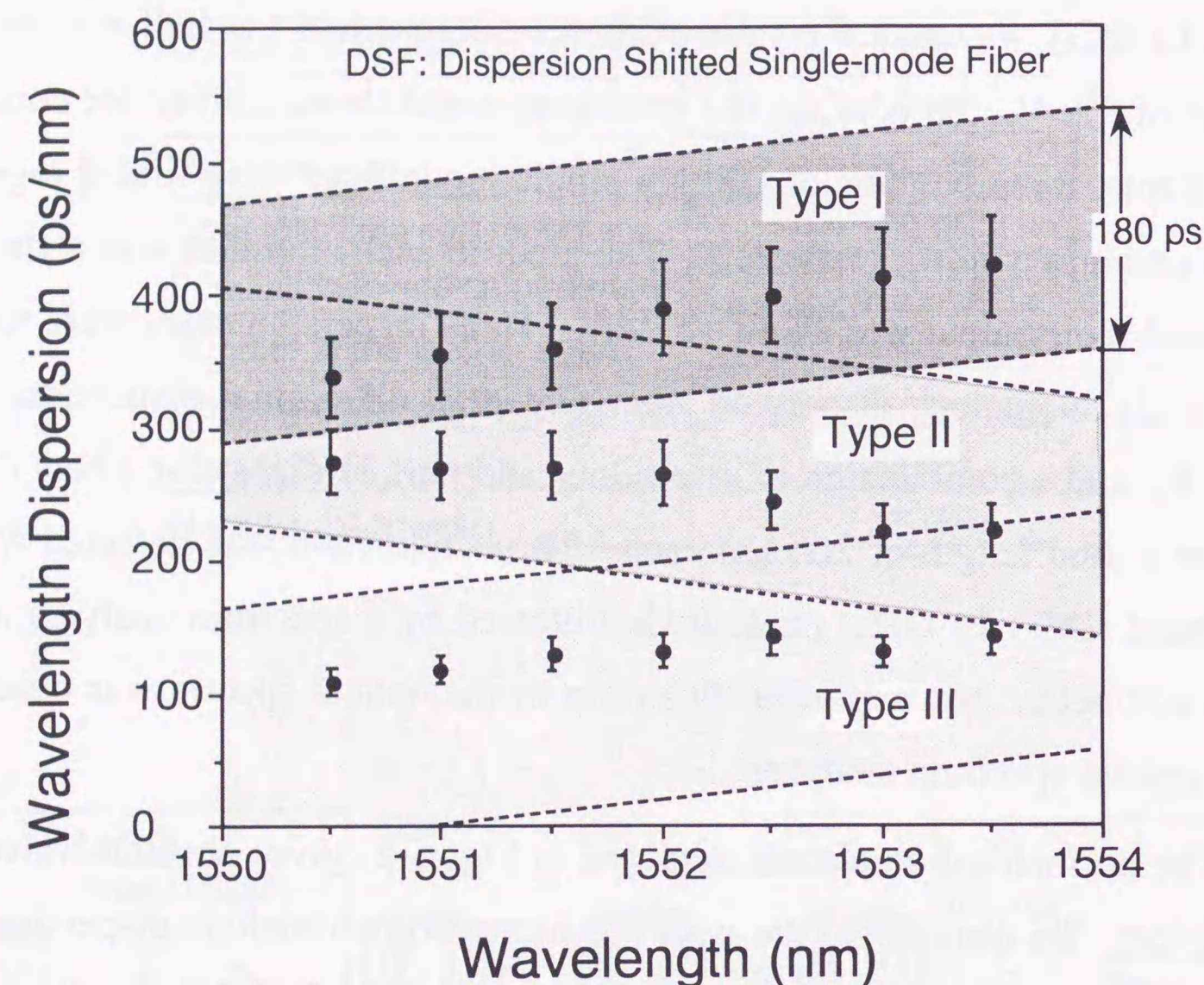


Fig. 6-6 Experimental results compared with the data from a method summing up the dispersion for individual fiber pieces. Each dispersion data was obtained by differentiating the fitting curve of group-delay data at several wavelengths.

6.4 Discussion

One distinct advantage in the proposed PM-AM conversion method is that no fitting calculation is needed to obtain the dispersion data. The measurement is direct and quick. Most other methods require a quadratic polynomial to fit the group delay, and the dispersion data is obtained by differentiating the fitted delay curve. As a result, they tend to take time. A version of the fitting method intended for narrow (4 nm) windows requires 10 min. to obtain the 36 data points [67].

The method proposed in this chapter needs no reference path. The two adjacent Bessel components originating from the optical phase modulation cause beating, and each spectrum acts as the reference to the other. Since they travel through the same fiber, the problems associated with long-distance measurement have been eliminated. The method is stable against external fluctuations. The obtained dispersion data indicate the value right at the source wavelength.

One disadvantage of this method is its inability to distinguish the polarity of the dispersion. However, the experiment showed that a slight (3 nm) sweeping of the wavelength was sufficient to distinguish the polarity of the test fiber. This sweep can be achieved by changing the temperature of a conventional DFB laser (typically about 0.9 nm/°C) by 23 °C.

In the proposed scheme, amplitude modulation appears as the light travels through the dispersive fiber, as well as decreasing in its average intensity. Therefore, problems with nonlinearity arises primarily when measuring a multi-fiber span in which optical amplifiers are used to recover the fiber loss at each junction. Most of the nonlinear effect will occur at the last amplifier's output because there may already have some amplitude modulation transformed from the phase modulation. The lightwave with some amplitude modulation is amplified to an intensity possibly sufficient to cause the nonlin-

ear effects. We estimated the SPM effect after traveling the effective length L_{eff} of the last fiber section by

$$\Delta\phi_{SPM} \cong \gamma L_{eff} \left[4E_0^2 J_0(b_m) J_1(b_m) \sin \frac{\beta''(k-1)L\omega_m^2}{2} \right]^2 \exp[-2\delta\omega\tau_1], \quad (6-22)$$

where k is the number of the fiber section and L is length of each fiber section, and the only first term of the infinite series of (6-10) is considered [67]. This expression approximates the effect since we can ignore any nonlinear effect after propagating

$$L_{eff} = \frac{1}{\alpha} \{1 - \exp(-\alpha L)\} \quad (6-23)$$

the effective length [67]. If 3 pieces of 100 km fiber are linked together, the nonlinear phase modulation is estimated to be 0.03 rad after traveling the effective length from the last amplifier's output. Since this nonlinear phase modulation is only 1/20 that of the phase modulator, we can ignore the SPM effect in this experiment.

6.5 Summary

The PM-AM conversion method can measure chromatic dispersion in any dispersion region. We show in this chapter that a sinusoidally phase-modulated single-mode laser can be successfully used for determining the chromatic dispersion of concatenated fiber sections. The limit of measurability is given by two factors. The factor of PM-AM conversion limits the amplitude that can be observed in the small dispersion region. This limit can, however, be released if the modulation frequency is tunable to the higher. The other limitation comes from the source linewidth. It was shown that a wide linewidth not only reduces the converted amplitude modulation, but also degrades the SNR

in the large dispersion region. As expected no optical nonlinear effect, such as SPM, was observed.

Appendix

We derive approximated correlation functions (6-17) and (6-18) from the last term of (6-16). Assuming the process of the random phase is stationary, we begin with shifting the area of integration by $\tau/2$. Therefore without loss of generality, the starting integral can be written

$$W(u) = \left\langle \int_{t+u}^{t+u+\tau} \dot{\phi}(\theta) d\theta - \int_t^{t+\tau} \dot{\phi}(\theta) d\theta \right\rangle. \quad (6-A-1)$$

Next, we divide the variable into two cases, $u < \tau$ and $u \geq \tau$.

Case 1: $u < \tau$.

The integration variables for those two definite integrals in (6-A-1) overlap each other from $t+u$ to $t+\tau$. These integrals can be equivalently rewritten as

$$W(u) = \left\langle \int_{t+\tau}^{t+u+\tau} \dot{\phi}(\theta) d\theta - \int_t^{t+u} \dot{\phi}(\theta) d\theta \right\rangle. \quad (6-A-2)$$

Using the characteristic functional of a stationary Gaussian process, (6-A-2) becomes

$$W(u) = \exp \left[- \left\{ \frac{1}{2} \int_{t+\tau}^{t+u+\tau} d\theta_1 \int_{t+\tau}^{t+u+\tau} d\theta_2 \langle \dot{\phi}(\theta_1) \dot{\phi}(\theta_2) \rangle + \frac{1}{2} \int_t^{t+u} d\theta_1 \int_t^{t+u} d\theta_2 \langle \dot{\phi}(\theta_1) \dot{\phi}(\theta_2) \rangle \right\} \right]. \quad (6-A-3)$$

By evoking (6-3) and (6-4), namely zero-mean process and $\langle \dot{\phi}(t) \dot{\phi}(t+\tau) \rangle = 2\delta\omega \delta(\tau)$

$$\begin{aligned} W(u) &= \exp \left[- \left\{ \frac{1}{2} \int_{t+\tau}^{t+u+\tau} d\theta_1 \int_{t+\tau}^{t+u+\tau} d\theta_2 \Theta(\theta_2 - \theta_1) + \frac{1}{2} \int_t^{t+u} d\theta_1 \int_t^{t+u} d\theta_2 \Theta(\theta_2 - \theta_1) \right\} \right] \\ &= \begin{cases} \exp[-2\delta\omega u] & \text{for } u > 0 \\ \exp[2\delta\omega u] & \text{for } u < 0. \end{cases} \\ &= \exp[-2\delta\omega |u|] \end{aligned} \quad (6-A-4)$$

where $\Theta(\theta_2 - \theta_1) = 2\delta\omega \cdot \delta(\theta_2 - \theta_1)$.

Case 2: $u \geq \tau$

Since integration variables for two integrals in (6-A-1) do not overlap each other, calculation goes straight.

$$\begin{aligned} W(u) &= \left\langle \int_{t+\tau}^{t+u+\tau} \dot{\phi}(\theta) d\theta - \int_t^{t+u} \dot{\phi}(\theta) d\theta \right\rangle \\ &= \exp \left[- \left\{ \frac{1}{2} \int_{t+\tau}^{t+u+\tau} d\theta_1 \int_{t+\tau}^{t+u+\tau} d\theta_2 \langle \dot{\phi}(\theta_1) \dot{\phi}(\theta_2) \rangle + \frac{1}{2} \int_t^{t+u} d\theta_1 \int_t^{t+u} d\theta_2 \langle \dot{\phi}(\theta_1) \dot{\phi}(\theta_2) \rangle \right\} \right] \\ &= \exp \left[- \left\{ \frac{1}{2} \int_{t+\tau}^{t+u+\tau} d\theta_1 \int_{t+\tau}^{t+u+\tau} d\theta_2 \cdot \Theta(\theta_2 - \theta_1) + \frac{1}{2} \int_t^{t+u} d\theta_1 \int_t^{t+u} d\theta_2 \cdot \Theta(\theta_2 - \theta_1) \right\} \right] \\ &= \exp[-2\delta\omega \tau], \end{aligned} \quad (6-A-5)$$

(6-A-4) and (6-A-5) are used as (6-18) and (6-17) in the main text, respectively, and where $\Theta(\theta_2 - \theta_1) = 2\delta\omega \cdot \delta(\theta_2 - \theta_1)$.

Chapter VII

Concluding Remarks

Chapter I briefly reviewed the historical background of optical digital transmission systems. It covered the technical progress seen from multimode to single-mode. It has happened in both fiber and semiconductor lasers, even though the former is almost limited in silica based fiber, and the latter developed on various material systems such as Si, GaAs, and InP. Network evolution from analog to digital, and then from asynchronous to synchronous multiplexing was also introduced. Transmission codings to guarantee transmission error performance were reviewed with some examples. Fundamental functions of the multiplexing line terminals were presented and discussed.

In subsequent chapters, terminal and multiplexing equipment for high-speed digital optical transmission systems intended for inter-office trunk line was studied. Topics covered included construction and realization of IC set

design, supervisory and control signal channel by C bit modulation of mB1C transmission code, 8-bit interleaved Hamming code for the AU-4 in the SDH networks, and dispersion measurement for the purpose of self-controlling the transmitter wavelength selection. The major results of these studies are summarized as follows:

1. By comparing WDM and two TDM schemes (bit interleaving and byte interleaving), we conclude that TDM with byte interleaving is best for the near future SDH network system. We have proposed an architecture for a line-terminating multiplexer (LTM) that realizes simple circuit configuration by multiplexing channels in integral multiples of eight. The key circuit is an 8-channel row-column exchanger (RCX), which we have implemented as a Si-Bipolar LSI. We have also shown that the STM-64 LTM can be successfully constructed with Bi-CMOS, Si-Bipolar, and GaAs MESFET ICs. Test results show that an STM-64 or 10 Gbit/s line-terminating multiplexer is feasible. (Chapter II)
2. A four part regenerator section termination LSI chip set has been designed, fabricated, and successfully tested. Its additional functions include frame alignment, some multiplex section termination functions, test pattern generation, and virtual error insertion. The Si-bipolar gate-array LSIs process the STM-64 signal in the byte (8 bit) parallel manner at 1.24 GHz. This result confirms feasibility of the line-terminating equipment for the STM-64 system. (Chapter III)
3. It has been shown that C bit modulation for the purpose of SV signal transmission is feasible in gigabit lightwave systems. The C bit, not traffic pulses, is modulated according to the SV signal. This technique's advantage is that the main sequence is independent of SV input. This makes it possible

to send a relatively high bit rate SV signal ($> \text{Mb/s}$) with reasonable circuit scale. The experimental survey proved that error-free transmission can be achieved beyond 40% of relative modulation depth, even when using a conventional optical repeater with no special care taken to ensure circuit linearity. (Chapter IV)

4. An 8-bit interleaved Hamming code for the AU-4 was proposed for SDH line terminating multiplexers. It features up to 1-byte error-correcting capability in AU-4 by performing (2370, 2358) shortened Hamming coding for each interleaved group of the AU-4 signal block. This general code offers the same correction performance regardless of the VC size or STM level by simply arraying the requisite numbers of coders/decoders in parallel. Nothing has to be changed in the configuration of the physical interface circuits for the coded frame. The experimental correction of ASE error in a 156 Mbit/s optical preamplifier has confirmed the FEC performance and feasibility. (Chapter V)

As a simple extension of this work, a co-worker and the author has reported a serial (18880, 18865) shortened hamming code [68]. Although the serial coding limits the correction performance to one bit per codeword, it requires smaller numbers of the check bit (12 bits) than the parallel code described in Chapter V (12 Bytes). The reported experiment features a simplified matrix method for a parallel FEC processing at lower clock rates, which enables lower power consumption and economical implementations. Those two hamming codes were applied to an STM-64, 550 km nongenerative transmission line equipped with 5 linear amplifier repeaters with successful correction performances as predicted by the theory under random-error assumption [69]. The test equipment utilizes circuits formed on CMOS field programmable gate arrays (FPGAs). It enables remote

installation by downloading the circuit connection pattern, without replacing the hardware. The flexible hardware architecture will enable error correcting function on demand basis, not only for purpose of the experimental convenience. Those results are summarized by the co-worker [70]

5. The PM-AM conversion method can measure chromatic dispersion in any dispersion region. We showed that a sinusoidally phase-modulated single-mode laser can be successfully used for determining the chromatic dispersion of concatenated fiber sections. The limit of measurability is given by two factors. The factor of PM-AM conversion limits the amplitude that can be observed in the small dispersion region. This limit can, however, be released if the modulation frequency is tunable to higher values. The other limitation comes from the source linewidth. It was shown that a wide linewidth not only reduces the converted amplitude modulation, but also degrades the SNR in the large dispersion region. As expected no optical nonlinear effect, such as SPM, was observed. (Chapter IV)

By the method described in Chapter IV, we can only obtain absolute dispersion values of single-mode fibers. However a scheme combined with a nonlinear effect, self-phase modulation (SPM), makes it possible to distinguish their sign or polarity, as reported by a co-worker and the author [71]. The principle employs a fact that depending upon the polarity the chirping due to SPM tends to emphasize or compress the PM-AM conversion effect. This gives a simple procedure for detecting the polarity, if the light source is so boosted by an optical amplifier that the output power can be high enough to generate the SPM effect in the fiber under test. The polarity can be obtained by observing whether the measured normalized amplitude, or PM-AM conversion efficiency, increases or decreases with increasing power. Another nonlinear influence by modulation instability has also been

reported by the same party [72]. The second nonlinear effect exaggerate the nonlinear effect mentioned above when the launched power exceeds 11 dBm.

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