

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

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BEFORE THE PATENT TRIAL AND APPEAL BOARD

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SOLiD, Inc.  
Petitioner

v.

CommScope Technologies LLC  
Patent Owner

U.S. Patent Nos. 7,639,982; 8,326,218; 8,577,286; 9,332,402

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**DECLARATION OF MARIA P. GARCIA**  
**UNDER 37 C.F.R. § 1.68**

I, Maria P. Garcia, do hereby declare:

1. I am making this declaration on behalf of SOLiD Inc. (“Petitioner”) in the matter of the *Inter Partes* Reviews of U.S. Patent Nos. 7,639,982 (“the ’982 patent”), 8,326,218 (“the ’218 patent”), 8,577,286 (“the ’286 patent”), and 9,332,402 (“the ’402 patent”) (collectively the “IPR matters”). Unless otherwise stated, the facts stated in this Declaration are based on my personal knowledge.
2. I am the Intellectual Property research librarian with the law firm of Jones Day. I received a Bachelor of Science in Chemical Engineering from University of Southern California (USC) in May 1989. I worked in the Circulation and Acquisition department at the Education and Information Science Library at USC from December 1983 to February 1989. I formerly worked at Lyon & Lyon, where I was responsible for all aspects of patent and trademark research. I am familiar with library procedures for receiving, cataloging, indexing, and shelving periodicals in libraries, such as those at the University of California, San Diego (UCSD), the Linda Hall Library, and the University of California, Los Angeles (UCLA).
3. Exhibit 1009 in the IPR matters is: Philip M. Wala, “A New Microcell Architecture Using Digital Optical Transport,” 1993 43rd IEEE Vehicular Technology Conference, 585 (1993) (“Wala”).

4. The Wala article is contained within the 1993 43rd IEEE Vehicular Technology Conference (the “IEEE Conference Book”), as evidenced by the table of contents of the IEEE Conference Book (page xvi). Appendix A to this declaration is a copy of the library catalog, machine readable cataloging (MARC) record, cover, and table of contents of the IEEE Conference Book from the UCSD Library, as well as the Wala article contained therein from the UCSD Library. I have compared the copy of the Wala article from the IEEE Conference Book with Exhibit 1009 in the IPR matters and attest that Exhibit 1009 is a true and correct copy.

5. Page 3 of Appendix A is a MARC record indicating on line 008 that a copy of the IEEE Conference Book was accessible on July 20, 1993. According to library procedure, this copy of the IEEE Conference Book, including the Wala article therein, would have been shelved and locatable by search within one to two days of this date. Page 6 of Appendix A is a checkout card for the IEEE Conference Book from the UCSD Library indicating that the IEEE Conference Book was checked out several times in the 1994-1997 time period. Further the front matter of the IEEE Conference Book on page 7 indicates that the conference occurred in New Jersey on May 18-20, 1993, and page 8 of Appendix A indicates a copyright date of 1993.

6. Exhibit 1011 in the IPR matters is: K. Ishida *et al.*, “A 10-GHz 8-b multiplexer/demultiplexer chip set for the SONET STS-192 system,” IEEE J. Solid-State Circuits, 1936 (1991) (“Ishida”).
7. The Ishida article is contained within Volume 26, Issue 12 of the IEEE Journal of Solid-State Circuits (the “IEEE Journal”), as evidenced by the table of contents of the IEEE Journal (page 17 of Appendix B). Appendix B to this declaration is a copy of the library catalog, MARC record, cover, and a partial table of contents of the IEEE Journal from the Linda Hall Library, as well as the Ishida article contained therein. I have compared the copy of the Ishida article from the IEEE Journal from the Linda Hall Library with Exhibit 1011 in the IPR matters and attest that Exhibit 1011 is a true and correct copy.
8. The stamp on page 4 of Appendix B indicates that copy of the IEEE Journal was received by the Linda Hall Library on November 27, 1991. According to library procedure, this copy of the IEEE Journal, including the Ishida article therein, would have been shelved and locatable by search within three to five days of its stamped received date. Page 4 of Appendix B indicates a journal date for Volume 26, Number 12 of the IEEE Journal of December 1991.
9. Exhibit 1012 in the IPR matters is: Zygmunt Haas, “A Mode-Filtering Scheme for Improvement of the Bandwidth-Distance Product in Multimode Fiber Systems,” J. Lightwave Tech., Vol. 11, No. 7, 1125 (1993) (“Haas”).



10. The Haas article is contained within Volume 11, Number 7 of the Journal of Lightwave Technology (the “Journal”), as evidenced by the table of contents of the Journal (page 6 of Appendix C). Appendix C to this declaration is a copy of the library catalog from UCLA. Appendix C also includes a cover and table of contents of the Journal from the UCSD Library, as well as the Haas article contained therein. I have compared the copy of the Haas article from the Journal with Exhibit 1012 in the IPR matters and attest that Exhibit 1012 is a true and correct copy.

11. Page 6 of Appendix C is a table of contents that indicates that copy of the Journal at the UCSD Library was received on October 19, 1993. According to library procedure, this copy of the Journal, including the Haas article therein, would have been shelved and locatable by search within three to five days of its stamped received date. Page 6 of Appendix C indicates a journal date for Volume 11, Number 7 of the Journal of July 1993. Page 1 of Appendix C indicates that a copy of the Journal was shelved at “SEL/8270” in 1993 at the UCLA Library.

**Declaration**

12. I declare that all statements made herein on my own knowledge are true and that all statements made on information and belief that are believed to be true, and further, that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under Section 1001 of Title 18 of the United States Code.

Maria P. Garcia

Maria P. Garcia

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*Personal Communication -  
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### **1993 IEEE 43rd Vehicular Technology Conference**

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# A New Microcell Architecture Using Digital Optical Transport

Philip M. Wala  
Waseca Technology, Inc.  
RR4 Box 145, P.O. Box 286  
Waseca, MN 56093 USA

**Abstract** - The performance of microcells which rely on analog modulation of an optical or microwave carrier is limited by the dynamic range of the analog transport. This paper describes a new microcell design which relies on high speed, analog-to-digital conversion to digitize the entire cellular band, enabling transparent RF transport to be accomplished over a digitally modulated optical fiber. Laboratory and field tests of an ADC Kentrox microcell which uses this technique are described. The application of this technology to the future design of an all-digital cellular base station is also discussed.

## I. INTRODUCTION

The trend towards a "personal communications" environment requires the ability to deliver sufficient coverage and capacity to provide truly ubiquitous service. Coverage must be provided in areas not served by macrocell antenna installations, including shadow areas, tunnels, and the interiors of high-rise office buildings. Capacity must be sufficient to accommodate high concentrations of users in convention centers and train stations. In addition, as cells continue to decrease in size and increase in number, the cost of real estate becomes significant, requiring that this coverage and capacity be delivered using equipment occupying a minimum amount of space. Finally, the plurality of potential modulation techniques, including FM, TDMA and CDMA, provides incentive for an infrastructure configuration that maintains flexibility for future system upgrades.

## II. LIMITATIONS OF TRADITIONAL SOLUTIONS

Solutions to the problem of delivering coverage and capacity fall into three general categories: cellular repeaters, distributed radio microcells, and centralized radio microcells [1].

Cellular repeaters increase coverage by re-radiating off-the-air signals. Because existing channels are re-transmitted, there is no capacity increase. Also, system degradation can occur due to the fact that noise and interference are re-radiated, along with the desired signals.

Distributed radio microcells or "mini-cells" are miniature versions of cell sites, incorporating a separate receiver and exciter for each channel served. While such a configuration can add both capacity and coverage, size and cost will increase proportionately to the number of channels. Such a system also lacks the ability to upgrade to new modulation techniques without replacing all of the transceiver modules.

Centralized radio microcells transport a broadband spectrum of combined radio frequency (RF) signals between remote antenna sites and banks of centrally located single-channel transceivers. This configuration has the advantage of transparency to modulation technique and the ability to serve multiple channels without any change to the remote equipment.

Most products of this type, however, suffer from the disadvantage that the transport is accomplished by the analog modulation of an optical or microwave carrier. This severely limits the dynamic range of the system for two reasons.

First of all, there is a tradeoff between signal-to-noise ratio and third-order intermodulation performance, which is a function of the modulation depth per channel [2], [3]. It can be difficult to maintain the optimum operating point, especially in a system where the number of active carriers can be constantly changing. Ultimately, in an optical system, the third-order intermodulation performance of the diode laser becomes the limiting factor [2], [4].



Secondly, the nature of analog modulation is such that the system dynamic range is destined to degrade in proportion to the attenuation on the transport link. In an optical system, this degradation is equal to twice the optical attenuation in dB [5], yielding an optical budget of as little as 10 dB [6] and maximum fiber distances on the order of 10 km [7].

### III. A NEW APPROACH: DIGITAL RF TRANSPORT

Many of the problems associated with analog modulation could be circumvented if it were possible to first digitize the information signal and then transport it digitally. Up until now that possibility has not been considered for the transport of the broadband "A" or "B" cellular spectrum because of the difficulty of digitizing the 12.5 MHz bandwidth with enough speed, resolution and linearity to accurately reproduce the dynamic range of signals found in the mobile environment. Recently, however, analog-to-digital converters with sufficient performance capabilities have become available [8], and have been successfully incorporated into a newly announced fiber-fed microcell product [9].

This digital microcell system consists of two main components: a donor or "host" site interface, and a remote microcell unit (Fig.1)

In the forward or "downlink" path, the host interface takes the combined channels from the cell site transmitters and digitizes the broadband signal at a sample rate of 30.72 MHz. An optical transmitter adds overhead and signaling information and creates a 552.96 Mbps serial data stream which is used to digitally modulate an optical carrier. At the remote site, an optical receiver demodulates the incoming optical carrier, and reconstructs the original 30.72 MHz parallel word, which is then converted back to RF by a digital-to-RF converter. At this point, the combined forward link channels have been reconstructed, and are ready to be linearly amplified for transmission.

In the reverse or "uplink" path, signals coming in through the antenna from mobiles or handhelds are filtered to remove out-of-band signals. The 12.5 MHz spectrum is then digitized in much the same way as the forward link spectrum, and a similar serial data stream is constructed for digital optical transport. At the host site, the optical carrier is demodulated, and the original RF signal is reconstructed and sent on to the individual channel receivers.

This particular configuration also incorporates the use of two different wavelength lasers and a wavelength division multiplexer (WDM) in order to enable bi-directional transport of forward and reverse path RF over a single fiber.

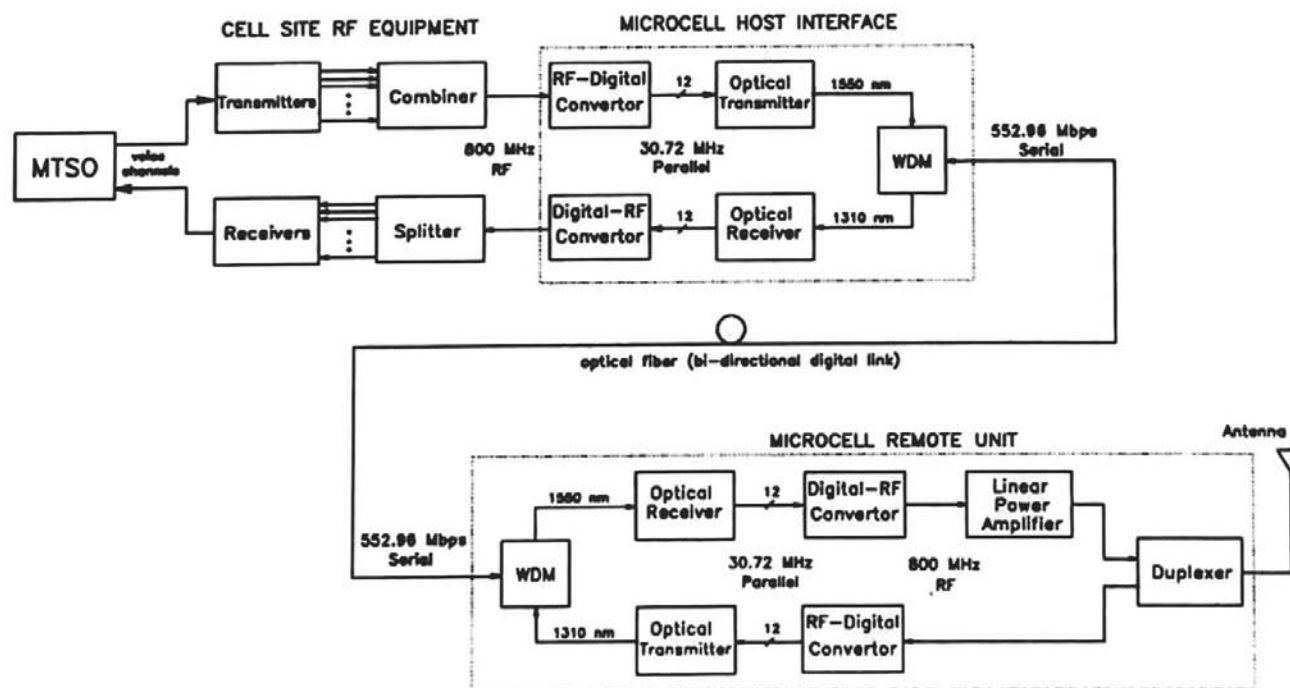


Fig.1. Digital RF transport microcell: system block diagram

The host interface for each microcell is contained in a 17" x 7" x 8.5" rack mountable module. The remote cabinet is 34" x 20" x 11" (including power supply and backup batteries), and can be pole or wall mounted.

#### IV. DIGITAL RF TRANSPORT: PERFORMANCE EVALUATION

Microcells employing the digital transport architecture have been tested on the bench as well as in actual service, and have demonstrated all the advantages of other centralized radio microcells, including transparency to modulation technique and channel selection, and a small remote unit size that is independent of the number of channels being carried. The absolute group delay due to the filtering and digitization process has been measured at less than 4  $\mu$ s across the band, and should not adversely affect TDMA or CDMA signals.

In addition, the digital microcell has proven to have some significant advantages over earlier generation analog versions. Because the forward link carries a relatively narrow dynamic range of signals, with the path performance typically limited by the linearity of the power amplifier (as in analog microcells), much of the following discussion will focus on the transport of the more hostile reverse link (mobile to cell site) spectrum.

Assuming that the digital transport can be accomplished without significant errors, the reverse link dynamic range will be limited by signal-to-noise ratio and linearity of the digitization and reconstruction process. In analog systems, the spurious free dynamic range is a function of the noise level in the signal bandwidth ( $N_s$ ) and the third-order intercept point ( $IP_3$ ), both in dBm:

$$SFDR = (2/3) \times (IP_3 - N_s) \quad (1)$$

Neglecting fiber loss, distributed feedback (DFB) lasers with an output power of +3 dBm have achieved a spurious free dynamic range on the order of 82 dB [4], with an absolute theoretical maximum of approximately 88 dB, assuming a 30 kHz signal bandwidth [10].

In the digitized system described here, (1) cannot be applied, because the third-order intermodulation products not only fail to rise faster than the fundamentals, their level may actually drop as the fundamental signals increase or additional signals are added (due, in part, to the "dithering effect", which can add effective bits of resolution to an over-sampled system). Consequently, one must evaluate the

dynamic range by noting the behavior of the noise floor and the third-order products as a function of signal level in order to determine the effect of any anomalies in the curves. In the prototype system, third-order spurious free dynamic range was limited to about 76 dB.

The real advantage of the digital transport method becomes apparent when one examines the optical budgets of the digital and analog systems. First of all, although the digital system third-order dynamic range is less than the theoretical maximum attributed to analog systems, it should be noted that this is achieved with a -4 dBm output laser, rather than the +3 dBm assumed earlier. An analog system with an equivalent laser power would be limited to a maximum dynamic range of about 74 dB.

Secondly, the nature of digital transport is such that, up until the point where the optical level approaches the sharp threshold where synchronization is lost, there is no degradation of the signal being transported whereas analog systems suffer 2 dB of degradation for every dB of optical attenuation [5]. The digital transport can maintain the RF path at its full dynamic range for optical levels as low as -31 dBm (more than a 20 dB improvement over analog transport). The dramatic nature of the improvement can be seen in Fig. 2, with an arbitrary 65 dB dynamic range performance requirement marked to show the effect on the usable fiber distance.

#### V. DIGITAL RF TRANSPORT: FUTURE APPLICATIONS

The dynamic range and optical budget advantages already demonstrated constitute only a portion of the potential benefits of digital RF transport.

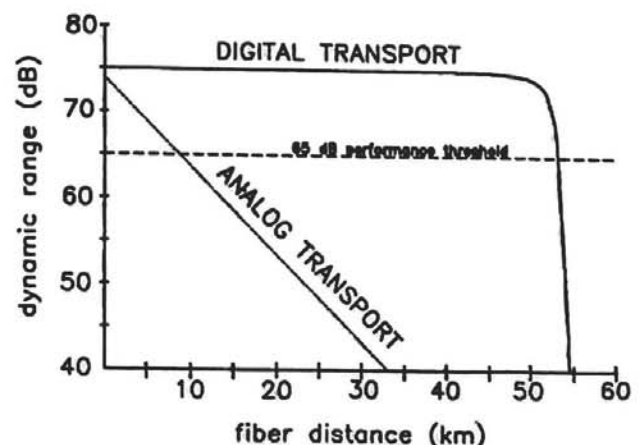


Fig. 2. Dynamic range vs. fiber distance (-4 dBm lasers)



Further potential advantages stem from the fact that digital signals are well suited to being time-division multiplexed with other digital signals. Although the current system uses time-division multiplexing to carry secondary digital channels with alarm and control information, secondary channels could serve a multitude of other purposes, including an uplink diversity path, an alternate RF service (such as paging), or digitized cable TV signals.

Ultimately, the establishment of a standard for the digital transport of RF could lead to a digital RF distribution system, carrying cellular or personal communications services, paging, trunking, or even broadcast signals, with drop and add capability, to a widespread network of microcells situated to provide seamless coverage of a service area.

Perhaps the biggest potential for the future of digital transport can be anticipated by another look at Fig.1. Because the voice channels between the mobile telephone switching office (MTSO) and the cell site are often already in digitized form, and the interface to the host microcell interface optical transceiver is a parallel digital word, one might legitimately wonder why one should even bother to make the intermediate conversion to RF. Conceivably, the cell site transceivers, combiners, splitters, and RF-digital converters could all be replaced by a digital signal processing unit, which would take the digitized voice signals and synthesize a digitized RF spectrum, reversing the process on the uplink path. In this configuration, RF would appear only at the remote microcell cabinet, and all cell site RF transceivers, combiners, splitters and filters would be replaced by an all-digital base station.

## VI. CONCLUSIONS

The application of new digital RF transport technology to a fiber-fed microcell application has been described, and its performance compared with that of analog microcells. Digital microcells have demonstrated dynamic range as good as or better than their analog counterparts, with no discernible degradation for fiber distances in excess of 50 km. Digital RF transport also has the potential for revolutionizing the way we think about distributing RF by making possible a digital RF transport infrastructure, in which a network of remote antenna locations are served, without degradation, from a centrally located all-digital base station.

## ACKNOWLEDGMENTS

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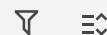
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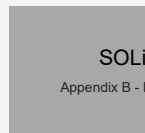
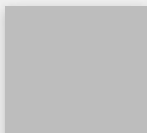
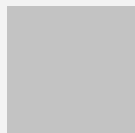
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# A 10-GHz 8-b Multiplexer/Demultiplexer Chip Set for the SONET STS-192 System

Kenji Ishida, Hirotsugu Wakimoto, Kunio Yoshihara, Mitsuo Konno, Shoichi Shimizu, Yoshiaki Kitaura, Kenichi Tomita, Takashi Suzuki, and Naotaka Uchitomi

**Abstract**—An ultrahigh-speed 8-b multiplexer (MUX) and demultiplexer (DMUX) chip set has been developed for the synchronous optical network (SONET) next-generation optical-fiber communication systems, which will require data bit rates of about 10 Gb/s. These IC's were designed using three novel concepts: a tree-type architecture giving reliable operation, a dynamic divider with a wide operating range, and a 50- $\Omega$  on-chip transmission line with high-speed pulse propagation. They were fabricated using a 0.5- $\mu\text{m}$   $\text{WN}_x$ -gate GaAs MESFET process. The DMUX and MUX operated up to 10.4 and 11.4 GHz, respectively, both with an adequate phase margin of more than  $230^\circ$ .

## I. INTRODUCTION

IN CURRENT optical-fiber communication systems, the data bit rate is 2.49 Gb/s. Higher data bit rates will be required for future increases in transmission capacity. An ultrahigh-speed data rate of 9.95 Gb/s is expected to be required for the optical-fiber communication systems associated with the synchronous optical network (SONET) STS-192 [1]. Fig. 1 shows an example of multiplexing and demultiplexing stages for the STS-192 system, which allows four-channel transmission with 2.49-Gb/s data rates. In this system, the 8-b multiplexer (MUX) and demultiplexer (DMUX) chip set is indispensable for byte interleaving. High-speed multibit (more than 8 b) MUX and DMUX chip sets have been reported [2]–[4]; however, a chip set with both 8-b signal processing and 10-GHz operation has not yet been reported.

This paper describes an ultrahigh-speed GaAs 8-b MUX and DMUX chip set as a key component of next-generation optical-fiber communication systems. This 8-b MUX and DMUX chip set, operating up to 10 GHz, features a tree-type architecture giving reliable operation, a dynamic divider with a wide operating range, and a newly developed 50- $\Omega$  on-chip transmission line (LGCL) with high-speed pulse propagation, in conjunction with an optimized 0.5- $\mu\text{m}$   $\text{WN}_x$ -gate GaAs MESFET process.

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K. Ishida, H. Wakimoto, Y. Kitaura, T. Suzuki, and N. Uchitomi are with the ULSI Research Center, Toshiba Corporation, 1, Komukai-Toshiba-cho, Saiwai-ku, Kawasaki 210, Japan.

K. Yoshihara, M. Konno, and S. Shimizu are with the Research and Development Center, Toshiba Corporation, 1, Komukai-Toshiba-cho, Saiwai-ku, Kawasaki 210, Japan.

K. Tomita is with the Tanagawa Factory, Toshiba Corporation, 1, Komukai-Toshiba-cho, Saiwai-ku, Kawasaki 210, Japan.

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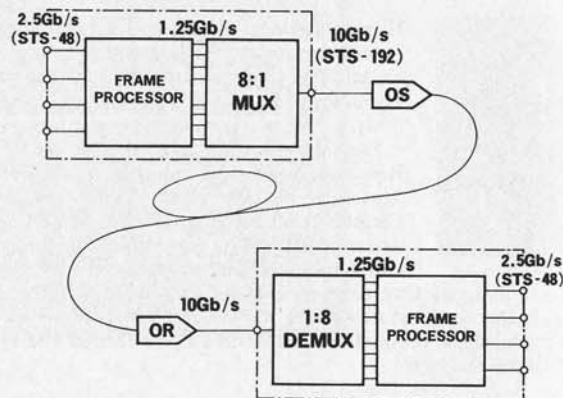


Fig. 1. Example of STS-192 system.

## II. CIRCUIT DESIGN

### A. Circuit Description

Both MUX and DMUX integrated circuits are implemented with the tree-type architecture and a dynamic divider in order to guarantee stable operation at 10 GHz, as shown in Figs. 2 and 3.

In Fig. 2, the 8-b MUX is realized with a 2-b MUX tree and a 1/8 counter with 1/2 dynamic divider front end. The 2-b MUX is composed of a phase-shifting D-type flip-flop (PS), a master-slave D-type flip-flop (MS), and a 2-b selector (SEL). The selectors for the first and second stages are integrated with the second- and third-stage MS and PS circuits (SMS and SPS). The SMS and SPS are constructed by using a three stacked gate structure [5]. The MS is paired with the PS in each 2-b MUX to increase the timing margin of the selector [6]. The eight-parallel input data streams have a data rate of 1.25 Gb/s. These data streams are multiplexed by 2-b MUX's in the first stage, resulting in four data parallel streams with data rates of 2.5 Gb/s. Similarly, the second and third stages use 2-b MUX's to generate a single output data stream with 10 Gb/s.

Fig. 3 shows a block diagram of the 8-b DMUX, with a 2-b DMUX tree, a 1/8 counter with dynamic divider front end, and a frame shifter. The 2-b DMUX is composed of the MS and the PS, and the PS is created by adding a slave stage to the MS in order to align the output signals. The frame shifter consists of five XOR gates and four selectors for the tree-type architecture.



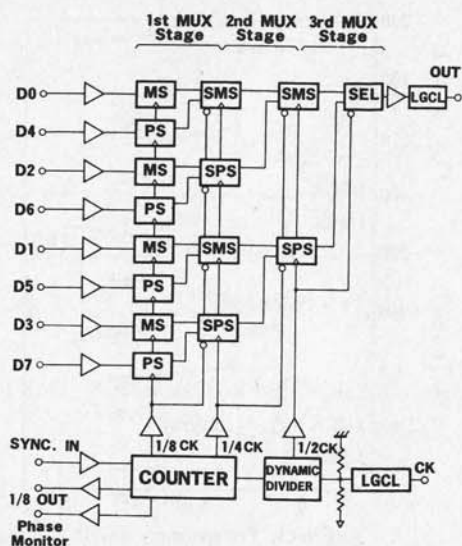


Fig. 2. Block diagram of 8-b MUX. MS: master-slave D flip-flop (DFF); PS: phase-shifting DFF; SMS: MS with 2-b selector; SPS: PS with 2-b selector; LGCL: ladder grounded coplanar line; SEL: 2-b selector.

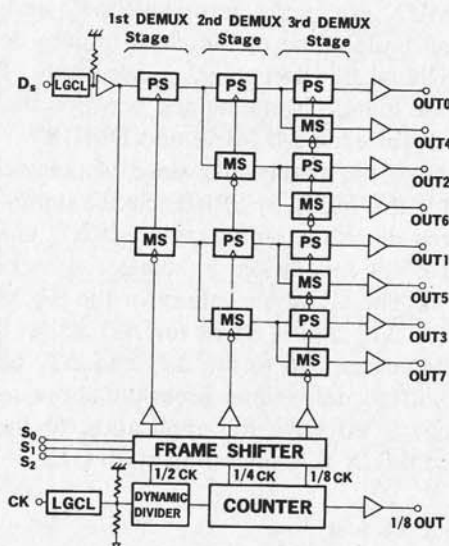


Fig. 3. Block diagram of 8-b DMUX.

The frame shifter rearranges the state of eight parallel output terminals (OUT0-OUT7) by changing a set of select signals ( $S_0$ ,  $S_1$ , and  $S_2$ ).

Both IC's use source-coupled FET logic (SCFL) to implement the master and slave latches as shown in Fig. 4. The logic swing voltage  $\Delta V$  and threshold voltage  $V_{th}$  were determined to be 0.9 and  $-0.2$  V, respectively, in order to maintain a noise margin larger than 0.35 V [5]. The FET gate widths in the flip-flops are determined to be 18, 10, and 4  $\mu\text{m}$  for the 1/2CK, 1/4CK, and 1/8CK flip-flops, respectively, so as to realize both high-speed operation and low power consumption. The final stage of the output buffer used an open drain for the MUX and a source follower for the DMUX. The MUX input and the DMUX output levels and the supply voltage were designed to be ECL compatible, that is, to drive a 50- $\Omega$  load

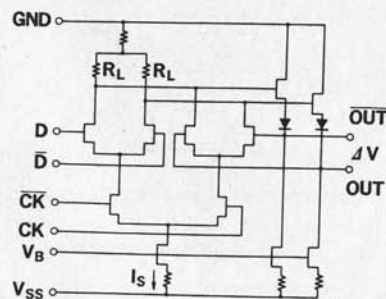


Fig. 4. SCFL circuit configuration.

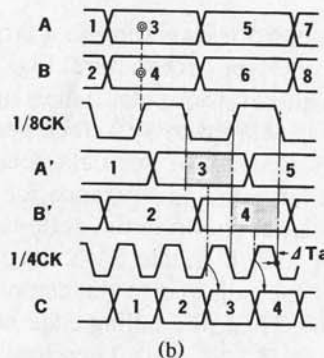
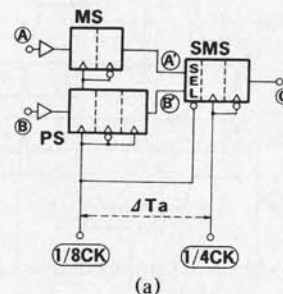


Fig. 5. Two-bit multiplexer: (a) block diagram, and (b) timing chart.

resistor at ECL logic levels ( $-1.3 \pm 0.45$  V) and operate on a  $-5.2 \pm 0.5$ -V supply voltage.

## B. 2-b MUX / DMUX Unit

Two problems must be taken into consideration when designing a MUX with tree-type architecture. First, precise control of the phase difference between the clock and input data signals for the 2-b selector is required. Here, a phase-shifting flip-flop (PS) was introduced to solve this problem [6]. The second problem is to ensure data latch because each stage flip-flop operates with different clocks. Thus, the timing relationships among three different clocks should be strictly determined to ensure reliable operation of the tree-type architecture.

Fig. 5(a) shows a block diagram of the 2-b MUX units. The selector is shown combined with the next-stage master-slave D-type flip-flop (SMS) as previously described. The timing relationships between clocks and signals must be taken into consideration to realize stable operation in this configuration. In particular, the delay time  $\Delta T_a$  between the 1/8CK and the 1/4CK clock signals is an

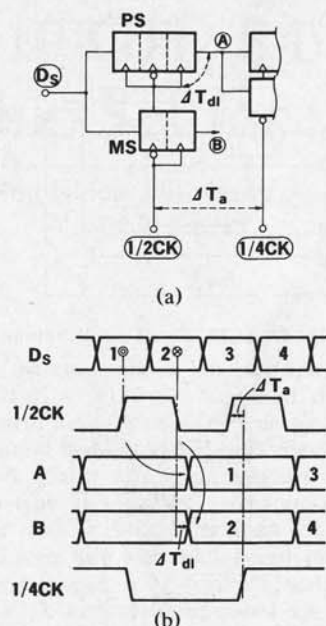


Fig. 6. Two-bit demultiplexer: (a) block diagram, and (b) timing chart.

important parameter for expansion to a larger number of channels in the tree-type architecture. Fig. 5(b) shows the timing relationship of two data signals and two clock signals. Data B is delayed by  $1/2$  clock period from data A due to the PS. Because of this data delay, the selector operates with a large timing tolerance for both negative and positive clocks, as shown in the dotted areas. In order to latch data A and B in the SMS flip-flop, the rising edges of  $1/4CK$  are adjusted to the center of the dotted areas in Fig. 5(b). Then, the falling edge of  $1/8CK$  is set to the edge of  $1/4CK$  ( $\Delta T_a = 0$ ). Therefore, this 2-b MUX with MS and PS D-type flip-flops has an important advantage over a conventional circuit without phase adjustment [7], [8].

Fig. 6(a) and (b) shows a block diagram and a timing diagram of the 2-b DMUX unit, respectively. The input data  $D_s$  are separated by the complementary pulse of  $1/2CK$ , and the output data A have the same phase as B due to the PS. Since the data A and B are delayed by  $\Delta T_{dl}$  compared with the falling edge of  $1/2CK$ , as shown in Fig. 6(b), the delay time  $\Delta T_a$  between  $1/2CK$  and  $1/4CK$  is made equal to  $\Delta T_{dl}$  in order to maintain a large phase margin for the next stage of the 2-b DMUX unit.

### C. Timing Relation of 8-b MUX / DMUX

The same timing relations can be applied to the 8-b MUX and DMUX. In the 8-b MUX and DMUX, there are three clock signals, as shown in Figs. 2 and 3, respectively. Hence, two interclock delay times ( $\Delta T_a$  and  $\Delta T_b$ ) must be adjusted in order to maintain a large phase margin, where  $\Delta T_a$  denotes delay time between  $1/8CK$  and  $1/4CK$  for the MUX and between  $1/2CK$  and  $1/4CK$  for the DMUX, and  $\Delta T_b$  is the delay time between  $1/4CK$  and  $1/2CK$  for the MUX and between  $1/4CK$  and  $1/8CK$  for the DMUX. Each timing delay is precisely

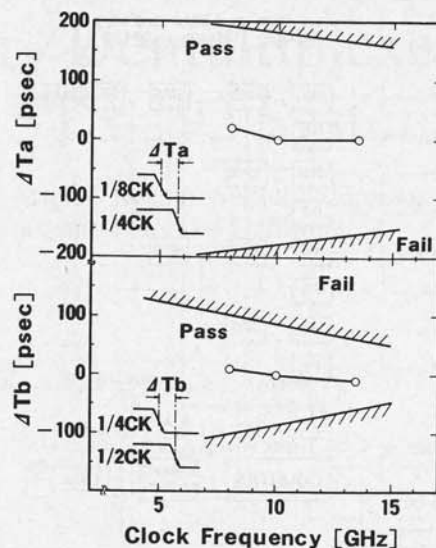


Fig. 7. Timing allowance of delay time,  $\Delta T_a$  and  $\Delta T_b$ , as a function of clock frequency.

adjusted using the delay of the buffer as follows:  $\Delta T_a$  and  $\Delta T_b$  for MUX are made zero, and  $\Delta T_a$  and  $\Delta T_b$  for DMUX are made equal to the clock-to-data delays  $\Delta T_{dl}$  in the individual flip-flop stages, respectively. Therefore, a large phase margin is maintained between the data and clock signals for each 2-b MUX and DMUX.

Fig. 7 shows the timing allowance of delay times  $\Delta T_a$  and  $\Delta T_b$  for the MUX by SPICE circuit simulation. The hatched area shows the failing region. If  $\Delta T_a$  and  $\Delta T_b$  are set to zero, the operation is stable, as mentioned in Section II-B. The allowable values for the 8-b MUX were  $\pm 180$  ps for  $\Delta T_a$  and  $\pm 80$  ps for  $\Delta T_b$  at 10 GHz. The open circles correspond to the  $\Delta T_a$  and  $\Delta T_b$  adjusted by using the buffer's delay time. From the above results, this methodology is effective for application to the multibit MUX and DMUX IC's operating at 10 GHz.

### D. Dynamic Divider

The high-speed counters in the MUX and DMUX IC's require an adequate frequency margin for stable operation at clock frequencies above 10 GHz. The dynamic divider, which meets the requirement for a 10-GHz input stage in the counter, is therefore used [9]–[11].

The fundamental circuit configuration of the  $1/2$  dynamic divider is shown in Fig. 8. SCFL circuits were used as buffer amplifiers because of supply voltage compatibility. This divider was designed for a wide operating range by optimizing the FET gate widths of the buffer amplifier and transfer gate. Furthermore, a source-follower stage is adopted in front of the output buffer to reduce load capacitance. Fig. 9 shows maximum operating frequency  $f_{max}$  as a function of the  $W_{gt}/W_{gb}$  ratio, where  $W_{gt}$  and  $W_{gb}$  are the FET widths for the transfer gate and the buffer amplifier, respectively. The parameter shows the number of gate fingers in the transfer gates and buffers, which have  $18\text{-}\mu\text{m}$  gate width as a basic width. The MESFET's have a threshold voltage of  $-0.2$  V and the



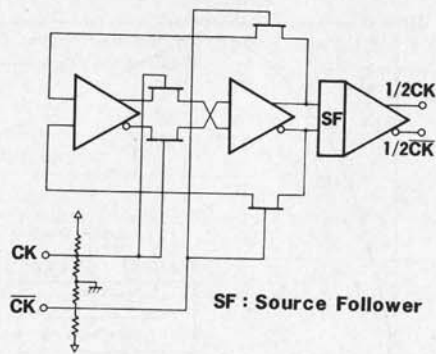


Fig. 8. Block diagram of 1/2 dynamic divider.

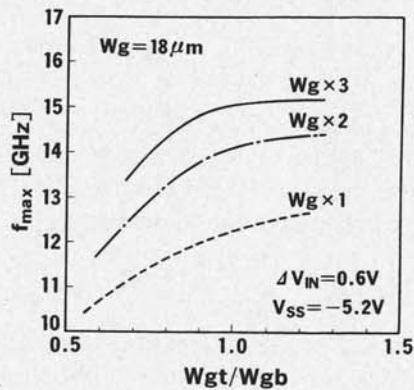
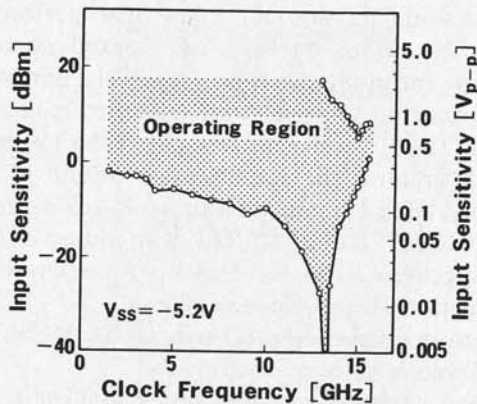
Fig. 9. Maximum operating frequency as a function of  $W_{gt}/W_{gb}$  ratio. The parameter is the number of buffer gate fingers.

Fig. 10. Input sensitivity as a function of clock frequency.

logic swing is 1.1 V. The input voltage amplitude and the gate width of the source-follower stage were maintained at 0.6 V and 10  $\mu\text{m}$ , respectively. In this figure,  $f_{\text{max}}$  increases with increasing  $W_{gt}/W_{gb}$  ratio. This is caused by the enhanced transconductance of the transfer gates. When the  $W_{gt}/W_{gb}$  ratio is larger than 1.0, the  $f_{\text{max}}$  shows a tendency to saturate. The phenomenon is more pronounced for the large buffer ( $W_g \times 3$ ). On the basis of the above results, the gate width of both the transfer gate and buffers was selected to be 54  $\mu\text{m}$  (18  $\mu\text{m} \times 3$ ).

Fig. 10 shows the measured data for the test device. This divider can operate from 1.5 to 15 GHz and has

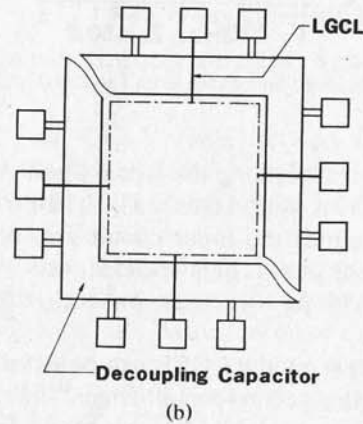
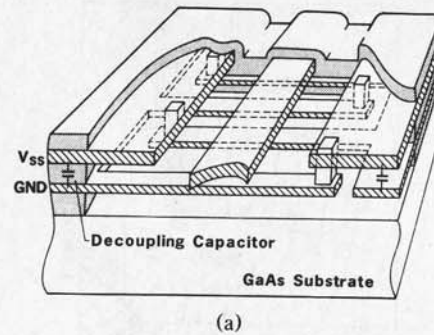


Fig. 11. (a) Ladder grounded coplanar line with decoupling capacitor. (b) Chip layout image for using LGCL.

sufficient input sensitivity for the initial 1/2 divider in the MUX and DMUX IC's.

### III. LAYOUT DESIGN

Impedance matching is becoming important for ultra-high-speed digital circuits as well as for analog circuits. In particular, there are three requirements for high-speed signal lines on chip: impedance matching, suppression of crosstalk, and power-line noise. The impedance of high-speed I/O ports must match those of the external systems in order to suppress pulse reflection and distortion. To solve these problems, a novel 50- $\Omega$  on-chip transmission line, named the ladder grounded coplanar line (LGCL), has been employed for the high-speed signal lines for the clock, MUX output, and DMUX input, as shown in Figs. 2 and 3. As for the DMUX input line and the clock (CK) lines for MUX and DMUX, on-chip termination resistors are placed at the other end of the LGCL in order to propagate the signals without the pulse reflection and distortion.

An LGCL is constructed of two segments with a microstrip line and a coplanar line. This LGCL can be quite freely designed and fabricated by using the two-layer interconnection without back metals. Fig. 11(a) shows the structure of the LGCL with a decoupling capacitor. The second-metal high-speed signal line is placed over the ladder-shape grounded first metal. The decoupling capacitor has a metal-insulator-metal structure and is laid down at both sides of the LGCL to prevent power-line

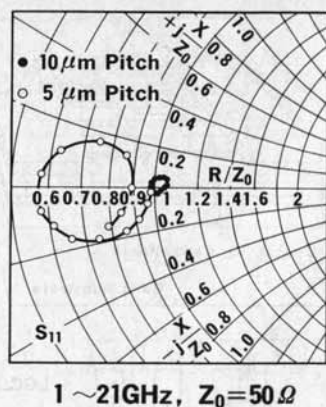


Fig. 12. Normalized impedance from 1 to 21 GHz ( $Z_0 = 50 \Omega$ ).

noise and crosstalk among the signal lines. A chip layout with this LGCL is illustrated in Fig. 11(b). A decoupling capacitor surrounds the inner circuit area to absorb the power-line noise due to high-speed signals. A decoupling capacitor of 150 pF was used on both the MUX and DMUX chips.

The impedance of the LGCL can be exactly controlled by changing the second-metal (signal line) width and first-metal (ground line) width and pitch. Fig. 12 shows the measured data for the LGCL impedance from 1 to 21 GHz. The first-metal and second-metal widths were fixed at 2.5 and 10  $\mu\text{m}$ , respectively, while the first-metal pitch was varied as a parameter. In 5- $\mu\text{m}$ -pitch LGCL (open circles), the impedance strayed from 50  $\Omega$  because of the large shunt capacitance. On the other hand, the 10- $\mu\text{m}$ -pitch LGCL (closed circles), which has the optimized capacitor, attained a constant impedance of 50  $\Omega$  up to 21 GHz. Therefore, the 10- $\mu\text{m}$ -pitch LGCL was employed as an on-chip signal line.

#### IV. FABRICATION PROCESS

##### A. 0.5- $\mu\text{m}$ FET Design

The MUX and DMUX were fabricated using 0.5- $\mu\text{m}$   $\text{WN}_x$ -gate GaAs buried p-layer LDD MESFET's with a shallow n-channel layer as shown in Fig. 13 (inset) [12].

In order to obtain maximum cutoff frequency at the target threshold voltage, it is important to determine the optimum ion-implantation energies and dosages for both the n-channel layer and buried p-layer. The buried p-layer dosage was determined by using a two-dimensional device simulator [13] with fixed Mg and Si ion-implantation energies. The Mg ion-implantation energy was fixed at 180 keV to surround the source/drain  $n^+$  regions and Si ion-implantation energy was fixed at 25 keV in order to form a shallow channel layer.

Fig. 13 shows the simulated result for the cutoff frequency  $f_T$  as a function of Mg dosage. The threshold voltage of  $-0.2$  V was kept constant through this simulation.  $f_T$  increases with increasing Mg dosage because of the enhanced transconductance due to the suppression of

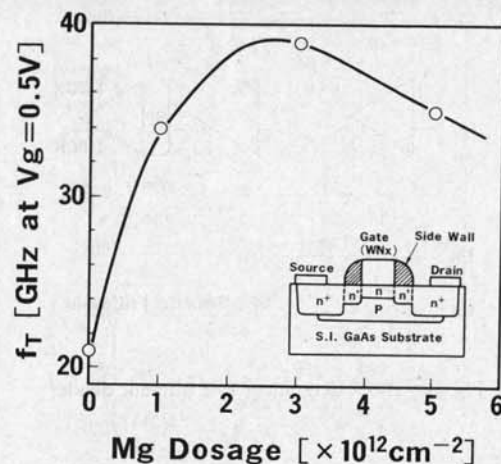


Fig. 13. Cutoff frequency  $f_T$  as a function of Mg dosage.

the short-channel effect. Further increases in Mg dosage to more than  $2.0 \times 10^{12} \text{ cm}^{-2}$  led to a reduction in  $f_T$ , because of the increase in gate-source capacitance and gate leakage current for large Si dosages.

##### B. Process Conditions

The fabrication process conditions for the MUX and DMUX chip set were as follows: the buried p-layer was formed by Mg ion implantation at 180 keV with a dose of  $2.0 \times 10^{12} \text{ cm}^{-2}$  as mentioned above. An n-channel layer was then formed by Si ion implantation at 25 keV with a dose of  $8.0 \times 10^{12} \text{ cm}^{-2}$  into undoped semi-insulating 3-in wafers. Post-implantation annealing was performed at 820°C for 20 min in an Ar + AsH<sub>3</sub> mixed atmosphere without any encapsulating film. The  $\text{WN}_x$  film was deposited by reactive RF magnetron sputtering in an Ar + N<sub>2</sub> mixed gas. The 0.5- $\mu\text{m}$  gate was patterned by modified optical lithography. The source/drain  $n'$  and  $n^+$  layers were formed by Si ion implantation at 45 keV with a dose of  $1.3 \times 10^{13} \text{ cm}^{-2}$  and at 110 keV with a dose of  $5 \times 10^{13} \text{ cm}^{-2}$ , respectively. The  $n^+$  region was separated from the gate metal by a 0.3- $\mu\text{m}$ -long sidewall.

The performance of the 0.5- $\mu\text{m}$  GaAs MESFET's is listed in Table I.

First- and second-interconnection metallizations were Ti/Pt/Au. The Au film thickness differed between the first (0.4  $\mu\text{m}$ ) and second (1.0  $\mu\text{m}$ ) levels, resulting in sheet resistivities of  $\rho_s = 0.07 \Omega/\square$  for the first level and  $\rho_s = 0.03 \Omega/\square$  for the second level. The interlayer insulating film and passivating film were 0.6- $\mu\text{m}$ -thick SiO<sub>2</sub> and 0.5- $\mu\text{m}$ -thick SiON, respectively. The first-metal insulator second-metal structure distributed capacitance was 0.1 fF/ $\mu\text{m}^2$ .

Both MUX and DMUX IC's were 3×3 mm<sup>2</sup>. Fig. 14 shows a photomicrograph of the DMUX IC chip, in which approximately 1800 circuit elements are integrated. The inner circuit is surrounded by the decoupling capacitor, composed of a metal-insulator-metal structure. The LGCL was used for DMUX input, MUX output, and clock signal lines.



TABLE I  
PERFORMANCE OF 0.5- $\mu\text{m}$  GaAs MESFET's

	Average	Standard Deviation
$L_g$ ( $\mu\text{m}$ )	0.50	0.02
$V_{th}$ (mV)	-160	40
$K$ (mS/V $\cdot$ mm)	424	18
$f_T$ (GHz)	40 (at $V_g = 0.4$ V, $V_d = 1$ V)	

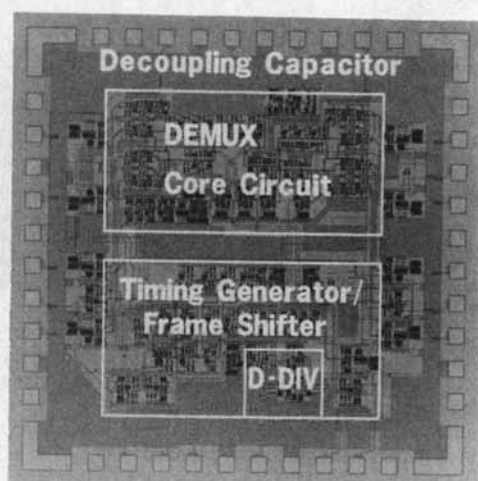


Fig. 14. DMUX photomicrograph.

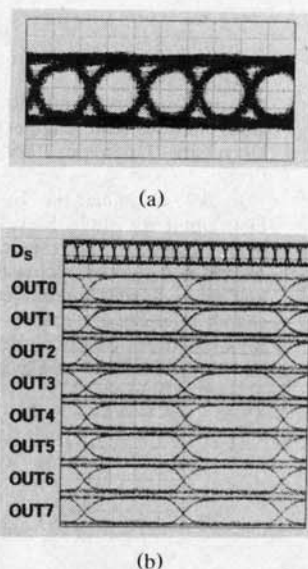


Fig. 15. Operating waveforms at 10 GHz: (a) MUX (0.2 V/div, 50 ps/div), and (b) DMUX (0.5 V/div, 200 ps/div).

## V. CHIP PERFORMANCE

High-speed device testing was performed directly on the wafer using a gigahertz-probe system. Fig. 15(a) shows the output waveforms of the MUX operating at 10 Gb/s. An error rate of less than  $10^{-10}$  was confirmed by using a pseudorandom signal with a word length of  $2^7-1$ . The IC power consumption was 3.8 W including the I/O internal

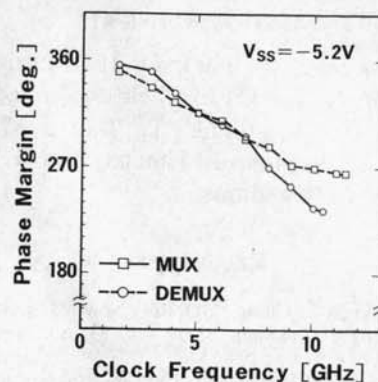


Fig. 16. Phase margin characteristics for MUX( $\square$ ) and DMUX( $\circ$ ).

circuit at a  $V_{ss}$  of -5.2 V. Note that very stable high and low levels in the output signal voltage were maintained. This is attributed to the effects of the LGCL and decoupling capacitor. The rise and fall times defined between 20 and 80% were 35 and 31 ps, respectively. Fig. 15(b) shows the input/output waveforms of the DMUX for a 10-Gb/s input data rate. Again, an error rate of less than  $10^{-10}$  was measured using a pseudorandom signal with a word length of  $2^{15}-1$ . The IC power consumption of the DMUX was 3.8 W.

Fig. 16 shows the phase margin between the data and the clock signals of the MUX and DMUX. The phase margin is an important characteristic of these high-speed devices used in this system. The MUX and DMUX operated from 1.5 GHz up to 11.4 and 10.4 GHz, respectively. These speeds were limited by the maximum frequencies of the measurement instruments. The phase margins of the MUX and DMUX are denoted by squares and circles, respectively. The phase margin of both MUX and DMUX is  $230^\circ$  even at 10 GHz. This result shows that these MUX and DMUX have phase margins that are sufficient for next-generation optical-fiber communication systems with an ultrahigh-speed data rate of 10 Gb/s.

## VI. CONCLUSION

An ultrahigh-speed GaAs 8-b MUX and DMUX chip set has been designed and fabricated with the following features:

- 1) a tree-type architecture which gives reliable operation;
- 2) a dynamic divider with a wide operating range;
- 3) a novel 50- $\Omega$  transmission line (the LGCL) for high-speed signal ports and on-chip decoupling capacitor; and
- 4) an optimized 0.5- $\mu\text{m}$  GaAs MESFET process.

Very stable operation at 10 GHz with a phase margin of more than  $230^\circ$  was realized, thus proving the applicability of these IC's to future optical-fiber communication systems (STS-192) as key devices.

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**Hirotsugu Wakimoto** received the B.Sc. and M.Sc. degrees in physics from Hokkaido University, Japan, in 1984 and 1986, respectively.

In 1986 he joined the Toshiba Research and Development Center, Kawasaki, Japan, where he has been engaged in the research and development of GaAs digital IC's. He is presently a Technical Staff Member at the Toshiba ULSI Research Center working on high-speed digital circuit design.

Mr. Wakimoto is a member of the Institute of Electronics, Information and Communication Engineers of Japan.



**Kunio Yoshihara** received the B.E. and M.E. degrees in electrical engineering from Kumamoto University, Japan, in 1978 and 1980, respectively.

In 1980 he joined the Toshiba Research and Development Center, Kawasaki, Japan. He was engaged in the research and development of hybrid integrated circuit technologies, such as a multichip package of X-ray computer tomography. Since 1984 he has engaged in the research and development of high-speed signal connections and high-speed digital circuit design.

Mr. Yoshihara is a member of the Institute of Electronics, Information and Communication Engineers of Japan.



**Mitsuo Konno** received the B.E. and M.E. degrees in electrical engineering from Tohoku University, Japan, in 1968 and 1970, respectively.

In 1970 he joined the Toshiba Research and Development Center, Kawasaki, Japan. He has been engaged in the research and development of MMIC's, such as microwave power amplifiers and phased-array antenna components. His research interests are mainly high-frequency and high-speed passive circuit design for GaAs IC's.

Mr. Konno is a member of the Institute of Electronics, Information and Communication Engineers of Japan.



**Kenji Ishida** received the B.E. and M.E. degrees in electronics engineering from the Toyohashi University of Technology, Japan, in 1981 and 1983, respectively.

In 1983 he joined the Toshiba Research and Development Center, Kawasaki, Japan, where he has been engaged in the research and development of GaAs devices. His research interests are mainly high-speed GaAs digital IC design. He is presently a Research Scientist at the Toshiba ULSI Research Center.

Mr. Ishida is a member of the Institute of Electronics, Information and Communication Engineers of Japan and the Japan Society of Applied Physics.



**Shoichi Shimizu** received the B.E. and M.E. degrees in electrical engineering from Yamanashi University, Japan, in 1970 and 1972, respectively.

In 1972 he joined the Toshiba Research and Development Center, Kawasaki, Japan. He has been engaged in the research and development of high-frequency analog IC's, high-speed digital IC's, and GaAs digital IC's. His research interests are mainly high-frequency analog circuit design and high-speed digital circuit design.

Mr. Shimizu is a member of the Institute of Electronics, Information and Communication Engineers of Japan.





**Yoshiaki Kitaura** received the B.E. degrees in metallurgical engineering from the National University of Yokohama in 1982.

In 1982 he joined the Toshiba Research and Development Center, Kawasaki, Japan. He has been engaged in the research and development of GaAs devices. His research interests are mainly in the process technology of self-aligned gate GaAs MESFET's. He is presently a Research Scientist at the Toshiba ULSI Research Center.

Mr. Kitaura is a member of the Japan Society of Applied Physics.



**Takashi Suzuki** graduated from Zushi Kaisei High School, Kanagawa, Japan, in 1969.

In 1982 he joined the Toshiba Research and Development Center, Kawasaki, Japan. He was engaged in the research and development of lithography technology. He is presently a Technical Staff Member at the Toshiba ULSI Research Center working on GaAs device technology.

Mr. Suzuki is a member of the Japan Society of Applied Physics.



**Naotaka Uchitomi** received the B.Sc. and M.Sc. degrees in chemistry from Kanazawa University, Japan, in 1976 and 1978, respectively. From 1978 to 1982 he studied physical chemistry at the Tokyo Institute of Technology.

In 1982 he joined the Toshiba Research and Development Center, Kawasaki, Japan, where he was engaged in the research and development of GaAs digital IC's, especially in the development of self-aligned-gate GaAs MESFET's with tungsten nitride gates. He is presently a Research Scientist at the Toshiba ULSI Research Center.

Dr. Uchitomi is a member of the Physical Society of Japan and the Japan Society of Applied Physics.



**Kenichi Tomita** received the B.E. degrees in metallurgical engineering from Kyoto University, Japan, in 1986.

In 1986 he joined the Toshiba Research and Development Center, Kawasaki, Japan, where he was engaged in the research and development of GaAs digital IC's. He is currently working on Si-bipolar process technology at the Toshiba Tamagawa Factory.

Mr. Tomita is a member of the Japan Society of Applied Physics.



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# A Mode-Filtering Scheme for Improvement of the Bandwidth-Distance Product in Multimode Fiber Systems

Zygmunt Haas, *Senior Member, IEEE*, and Mario A. Santoro, *Member, IEEE*

**Abstract**—In this paper, we investigate a scheme in which the *bandwidth-distance* product of a multimode fiber is extended, so that it can nearly support the transmission rate of single-mode systems. Some pioneering work suggest that limited mode launch may improve the system performance by reducing the modal dispersion and increasing the system bandwidth. The scheme investigated in this paper is based on selective launching of lower order modes into the fiber at the transmitting end and on filtering out at the receiver the fraction of the energy that was coupled into the higher order modes throughout the propagation in the multimode fiber. We investigate here the power penalty and the dispersion performance of the scheme. In particular, we show that the scheme carries about 6.5-dB penalty and we demonstrate doubling of the *bandwidth × distance* value. Moreover, the effect of splices on the scheme performance is also presented. We envision that the scheme may be used for upgrading the existing multimode systems without the need to rewire these installations. For example, the scheme can be used to replace the FDDI installations with future high-speed networks, when transmission rates increase from megabits per second to gigabits per second.

## I. INTRODUCTION AND MOTIVATION

THE motivation behind this work is to propose a scheme that will enable an easy (i.e., economical) upgrade of a communication system based on optical multimode fiber from a network operating at tens to hundreds of megabits per second, such as fiber distribution data interface (FDDI), to a network capable of supporting gigabit per second transmission. On one hand, the relatively large amount of multimode fibers already deployed in buildings for FDDI and similar LAN's will not, presently, support such an increase in speed to gigabits per second. On the other hand, the recent increased interest in very-high-speed networks, may shortly turn the gigabit per second communication networks into reality. Thus a technique that will enable reuse of the multimode fibers in high-speed transmission is of particular importance. Moreover, some system designs call for the use of multimode fibers, such as some military installations. Our scheme can be used in these cases to support increased transmission rates.

Multimode fibers were widespread when the first optical systems were designed, because it was easier and cheaper to work with a multimode fiber than with a single-mode fiber. In fact, some standards call for multimode fibers (FDDI).

However, the *bandwidth-distance* product of the multimode fiber is much smaller than the *bandwidth-distance* product of a single-mode fiber. Thus when increase in the transmission rate in a multimode system is needed, it may be required to rewire the whole system with single-mode fiber. This may be, of course, quite an expensive proposition in many circumstances.

The potential of multimode fibers for high-bandwidth transmission, as well as evaluation of the performance of multimode fiber dispersion was described in several papers. Mode-dependent characteristics, such as attenuation, for example, were explored by various researchers (e.g., [1]). Impulse response and the obtainable data rates of multimode fibers investigated in [2] indicating that mode coupling may, in fact, reduce the effect of modal dispersion.<sup>1</sup> In [3], a study of lower order mode launching is described for the purpose of learning the relative contribution of wall imperfections and bends to the mode-mixing phenomenon. The study concludes that the amount of mode conversion due to wall imperfection is negligible and that it is expected that single-mode operation may be possible over considerable distances. A study of multimode liquid-core fibers in [4] shows that pulse dispersion depends on launched mode distribution and on mode conversion, which is primarily a result of the bend radius. Furthermore, [4] suggested that mode filtering by selectively absorbing higher order modes can be used to maintain wide bandwidth. Similarly, [5] reports on the dependence of pulse broadening in high-loss multimode fibers, concluding that axial launching into the core is essential for minimum dispersion. Several authors dealt with analytical estimation of system response. For example, [6] proposes a bundle ray model with angular distribution of power characteristic to Gaussian beam and [7] models the response of a concatenated system for the low- and the high-order modes. It is suggested in [7] that the system performance depends on the actual distribution of power among the propagated modes.

Based on the above studies, we propose here a practical scheme, in which the *bandwidth-distance* product of a multimode fiber can be easily extended to support nearly the same transmission rate as the single-mode system. The performance of the scheme are investigated.

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The authors are with AT&T Bell Laboratories, Holmdel, NJ 07733.  
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<sup>1</sup> It is pointed out in [2] that the loss associated with the coupling may limit the usefulness of such a scheme.

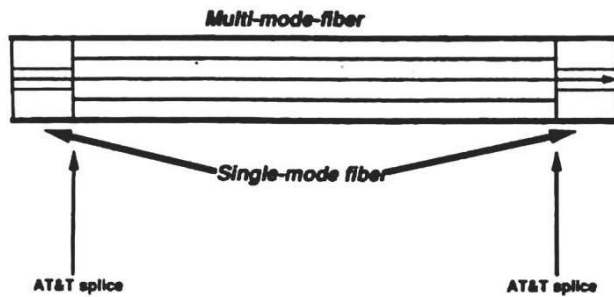


Fig. 1. Principle of the proposed scheme for extension of the bandwidth-distance product in multimode fibers.

## II. EXTENDING THE BANDWIDTH-DISTANCE PRODUCT FOR MULTIMODE FIBERS

Since modal dispersion in a direct detection system is a multipath, nonlinear phenomenon, it can be partially corrected by nonlinear post-detection adaptive filtering. We, however, propose here a scheme that operates in the optical domain. The idea behind the scheme is to reduce the number of propagated modes by launching a limited number of lower order modes into the multimode fiber and by recovering a limited number of modes on reception. The principle of the scheme is shown in Fig. 1, where we show ideal propagation in a single-mode fiber. By exciting a limited number of lower order modes, only a small portion of the total power will be coupled into the other modes. And since only the lower order modes are retrieved at the receiving end, the power contributing to the modal dispersion is considerably reduced. The residual mode dispersion is created by two elements: the modal dispersion among the lower order modes, and the power coupled from the lower order modes into the higher order modes and back into the lower order modes. This second element is relatively small, as suggested in [8]. Consequently, the expected penalty of the filtered systems as opposed to a regular single-mode system is relatively low. We investigate this penalty in this paper.

The basic experimental setup is shown in Fig. 2. 2.3 km of multimode fiber is placed on a spool.<sup>2</sup> The fiber was manufactured by AT&T with the following parameters: 62.87  $\mu\text{m}$ /125.24  $\mu\text{m}$  core/cladding diameters, attenuation of 0.86 dB/km at 1300  $\mu\text{m}$  and bandwidth  $\times$  distance of 1114 MHz at 1300  $\mu\text{m}$ .<sup>3</sup> The laser is ASTROTEC 1216D. The receiver is ASTROTEC 1306AA, with capability of up to 1.7 Gb/s.

The laser terminates with a single-mode fiber (9- $\mu\text{m}$  core) pigtail. An attenuator was placed after the laser and an additional 1.6 m of single-mode jumper connects between the multimode fiber spool and the attenuator. This connection, which performs the initial filtering by launching only the lower order modes into the multimode fiber, is made with an AT&T rotary splice. These lower order modes propagate through the multimode fiber with relatively small coupling of power into the other modes. At the receiver side, the multimode fiber

<sup>2</sup>The FDDI standard defines 2 km as the maximal distance between adjacent nodes on the ring.

<sup>3</sup>This is rather high bandwidth  $\times$  distance value; four other samples are between 577 and 748 MHz  $\cdot$  km. We chose the 1114 MHz  $\cdot$  km fiber to demonstrate the "worst case" improvement. And still with this relatively high bandwidth-distance product, our scheme demonstrates quite significant improvement.

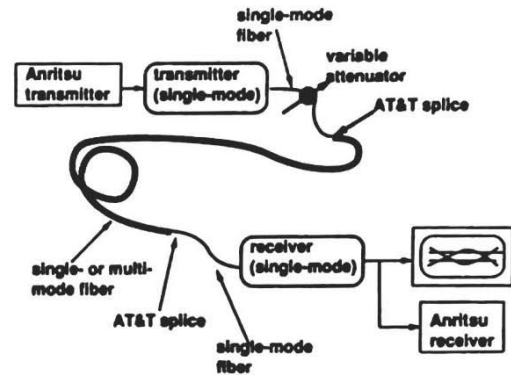


Fig. 2. The experimental setup.

is coupled by another AT&T rotary splice to a single-mode jumper of length 1.6 m, which is, in turn connected to the receiver pigtail. This splice at the receiver side provides the filtering means that removes higher order modes, which were excited by fiber-mode coupling from the initially launched modes.

Both of the elements of the proposed technique are essential: launching of the lower order modes and retrieving of the lower order modes. Launching of the lower order modes has two important implications: it concentrates most of the energy near the center of the fiber core, from where the energy is filtered at the receiving end, and reduces the amount of energy in the higher order modes, thus reducing the effect of modal dispersion from the coupling of the higher order modes into the lower ones.

In order to learn how the spatial distribution of the launched signal changes with propagation through the fiber, we looked at the near field emerging from a multi mode fiber in three different configurations. The near field was observed by focusing the output of multimode fiber on a CCD camera. First, we observed the near field of a multimode fiber with large number of modes excited (Fig. 3). This gives us the reference to calculate the spatial distribution of other cases (since the diameter of the multimode fiber is known to be 62.5  $\mu\text{m}$ ). The near-field distribution of a short piece of multimode fiber after launching the lower order modes is displayed in Fig. 4. The size of the beam is approximately 9  $\mu\text{m}$ , which corresponds to the size of the launched beam coming from single-mode fiber. Fig. 5 shows the size of the 9- $\mu\text{m}$  beam after propagation through 2 km of multimode fiber. The beam size appears to be  $\approx 15 \mu\text{m}$ , or only 1.7 times larger than the launched beam. These figures demonstrate that the beam remains more or less contained to the center of the core after passing the 2-km distance.

## III. PERFORMANCE OF THE SCHEME

Two performance criteria have to be considered here: signal attenuation and residual modal dispersion.

Signal attenuation is created by the filtering operation. The splice at the transmitter side introduces little loss of the optical signal, since all the energy in the single-mode fiber is concentrated in the lower order modes that is fully coupled into the multimode fiber. This is not the case at the receiving splice, where the filtering of the higher order modes removes

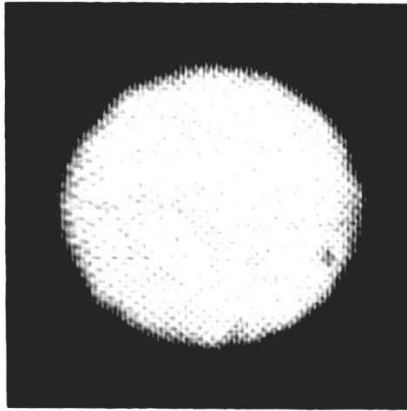
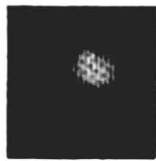
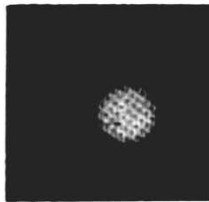


Fig. 3. Near-field distribution of a multimode fiber with all modes excited.

Fig. 4. Near-field distribution of a short piece of MM fiber coupled to 9- $\mu\text{m}$  beam.Fig. 5. Near-field distribution of 2 km of MM fiber coupled to 9- $\mu\text{m}$  beam.

some of the signal energy and thus results in some loss. The amount of this loss is difficult to calculate, since it is created by the mode-coupling phenomenon in the multimode fiber, where the spatial distribution of energy is not uniform. Moreover, this loss increases with distance, since the amount of coupling increases with distance. Some references quote quite a large loss (about 19 dB) at the multimode-to-single-mode junction ([9]). However, these references assume uniform distribution of energy or excitation of many modes in the multimode fiber. Obviously, this is not the case here. In order to evaluate the actual loss in our case, an experiment, shown in Fig. 6 was conducted. By varying the laser current supply, the power launched into the fiber under test was varied. The results are summarized in the following table ( $P_1$  and  $P_2$  are defined in Fig. 6):

$P_1$ [dBm]	$P_2$ [dBm]	Loss [dB]
-25.0	-30.40	-5.40
-20.0	-24.92	-4.92
-15.0	-20.56	-5.56
-10.0	-15.28	-5.28
-5.0	-9.69	-4.69
-0.0	-5.43	-5.43
+3.0	-2.10	-5.10

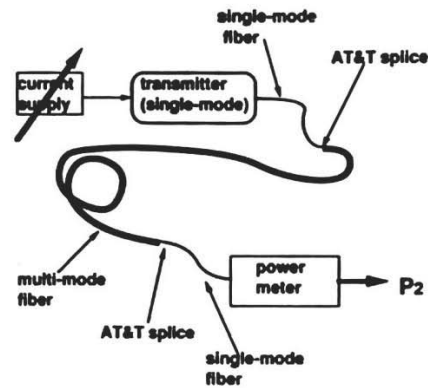
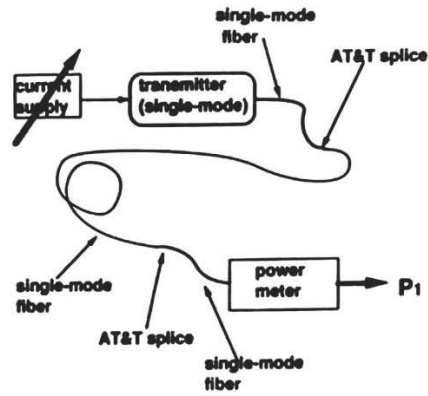


Fig. 6. Measuring the power and the loss of the filtering technique.

As evident from the above table, the excess loss for the above setup is on the order of 5 dB. This 5 dB is in good agreement with the observed beam divergence in Figs. 4 and 5. From these figures, a launched beam of 9  $\mu\text{m}$  diverged to a beam of 15  $\mu\text{m}$  in 2 km of multimode fiber. Coupling the 15- $\mu\text{m}$  beam into 9- $\mu\text{m}$  single-mode fiber results in a loss equal to square of the ratio of the two beams; i.e., 4.4 dB.

Besides the power loss, the scheme may still have some residual dispersion, since some of the power from the lower order modes is coupled into the higher order modes and back into the lower order modes. We evaluate this dispersion by comparison of the BER performance of the filtered multimode transmission with the single-mode system. In order to do so, we measured in both of the systems the BER as a function of bit rate with received power as a parameter. The tests were done using the a Bit-Error Rate tester and  $2^{10}-1$  PN sequence. The results are shown in Fig. 7. In Fig. 8, the BER is shown as a function of received power, for a selected number of bit rates. As a point of reference, the BER of the filtered system at 1 Gb/s and -30 dBm is about  $10^{-9}$ .

In Figs. 9 and 10, the performance of the same setup but with single-mode fiber instead of the multimode fiber is shown. At 1 Gb/s, the required received power for  $10^{-9}$  BER is -31.5 dBm; i.e., about 1.5-dB penalty in the filtered system. Figs. 11 and 12 superimpose the performance of the filtered and the single-mode systems. For large bit rate and small received power, the two curves are close to each other, indicating that the performance of the system is limited by the multimode

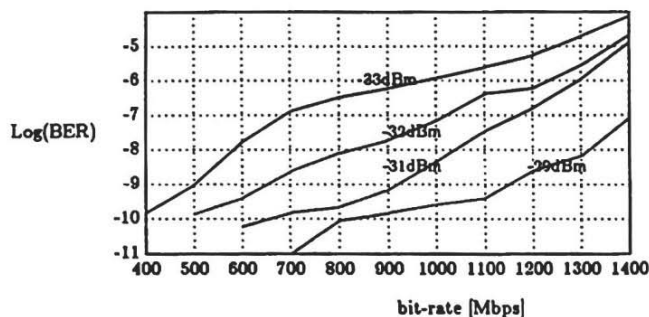


Fig. 7. Performance of the filtered system: BER versus bit rate.

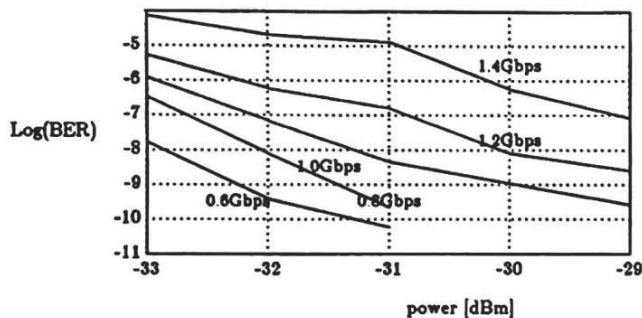


Fig. 8. Performance of the filtered system: BER versus received power.

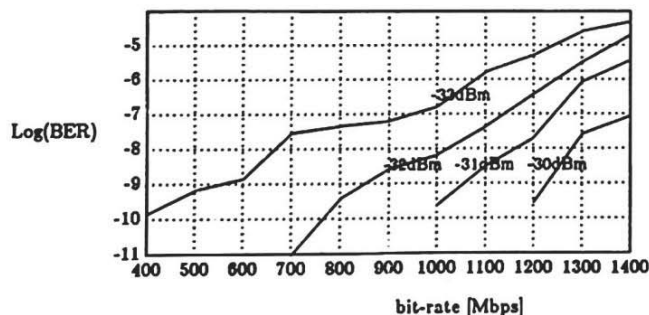


Fig. 9. Performance of the single-mode system: BER versus bit rate.

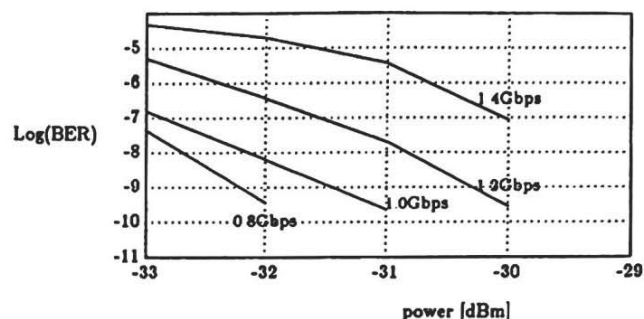


Fig. 10. Performance of the single-mode system: BER versus received power.

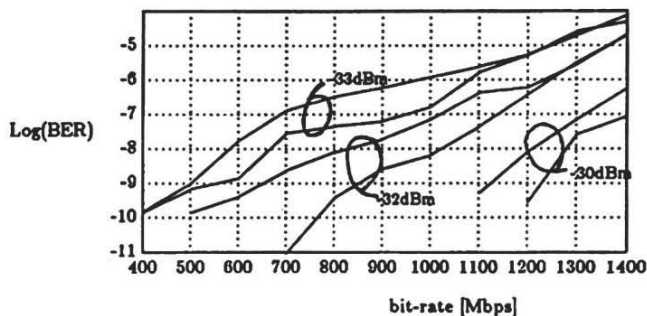


Fig. 11. Comparison of the systems: BER versus bit rate.

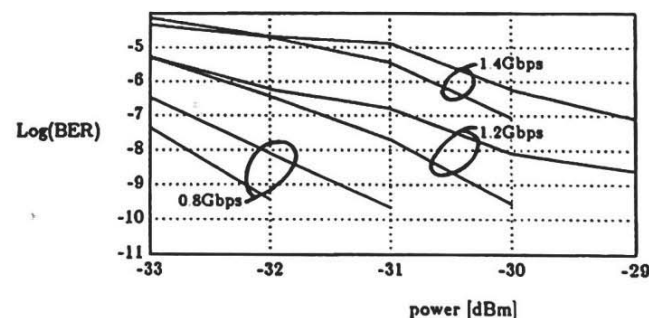


Fig. 12. Comparison of the systems: BER versus received power.

dispersion.<sup>4</sup> In the dispersion-limited regions, the penalty is between 1.0 to 1.5 dB for the filtering system. Thus coupled with the loss penalty due to the filtering operation of 5 dB, the total system penalty is about 6.5 dB. The *bandwidth × distance* product for the filtered system is 2185 MHz · km, and is calculated from Figs. 7, 9, and 12 by noting that the penalty of the system is about 1 dB at 950 MHz. Thus the improvement is nearly double the *bandwidth × distance* product; i.e., from 1114 to 2185 MHz · km. Even more dramatic improvement is achieved with multimode fiber with lower *bandwidth × distance*, as was the case with the other multimode samples (see footnote 2).

Next, we demonstrate the improvement in dispersion by the use of the mode filtering technique. In order to do so, the following set of experiments was performed:

1. Multimode launching, Multimode fiber, Multimode receiver (MM-MM-MM);

<sup>4</sup> In fact, the performance in these regions is limited by the receiver, whose sensitivity for  $3 \times 10^{-11}$  is -32 dBm at 1.7 Gb/s.

2. Multimode launching, Multimode fiber, Single-mode receiver (MM-MM-SM);
3. Single-mode launching, Multimode fiber, Multimode receiver (SM-MM-MM);
4. Single-mode launching, Multimode fiber, Single-mode receiver (SM-MM-SM)—this is the proposed scheme;
5. Single-mode launching, Single-mode fiber, Single-mode receiver (SM-SM-SM).

To perform these experiments, a mode scrambler (Model FM-1 Newport Corp.) was placed on the multimode fiber and an additional receiver with multimode pigtail was employed. The modified setup is shown in Fig. 13. The mode scrambler simulates the mode mixing as if a multimode fiber pigtail were coupled to a laser.<sup>5</sup> The multimode pigtailed receiver couples all the received modes to the detector. The resulting waveforms were recorded in Figs. 15–19. All the experiments

<sup>5</sup> We assume that the mode scrambling simulated in this way is sufficient for the purpose of our comparison. Appendix C, Page 1026



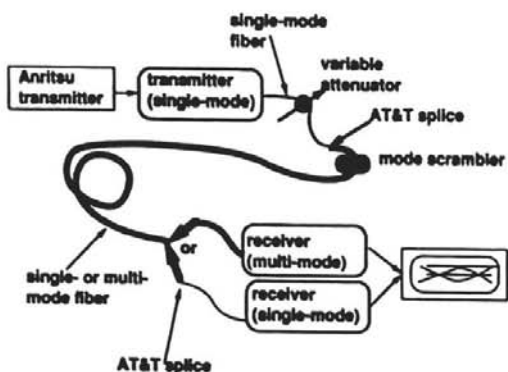


Fig. 13. The modified experimental setup.

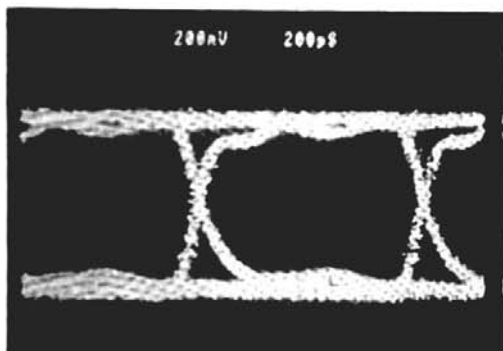


Fig. 14. Eye diagram for the direct connection.

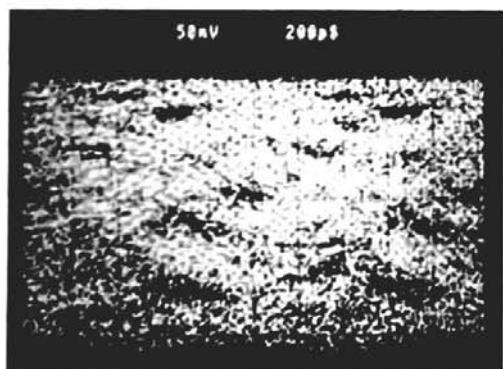


Fig. 15. Eye diagram for the MM-MM-MM experiment.

were performed at 1 Gb/s using the  $2^{10} - 1$  PN sequence, with a 2.3-km multimode fiber.

The eye diagram in Fig. 14 is the direct connection of the transmitter to the scope; i.e., without the optical link. Fig. 15 shows the amount of intersymbol interference for full multimode system operating at 1 Gb/s. Obviously, errorless reception cannot be achieved without any post-detection processing, since the eye is completely closed. The reception is significantly improved when a mode filter is placed either at the transmitting end (Fig. 16) or at the receiving end (Fig. 17). When both the changes are made simultaneously (filtering at the input and at the output), the performance is close to that of the full single-mode fiber system, as evidence from Figs. 18 and 19 (the fully single-mode system).

Finally, we investigate in Fig. 20 the effect of the splices on the performance of the proposed scheme. In this figure, the BER is shown as a function of the received power with

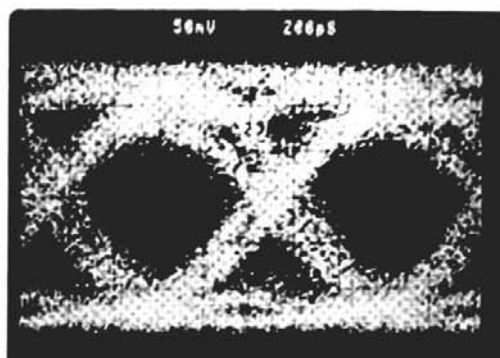


Fig. 16. Eye diagram for the SM-MM-MM experiment.

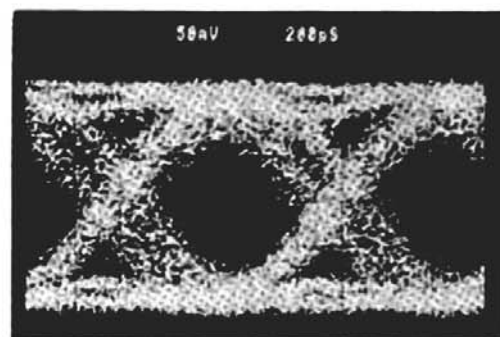


Fig. 17. Eye diagram for the MM-MM-SM experiment.

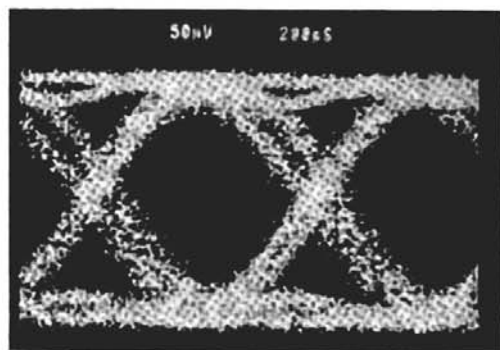


Fig. 18. Eye diagram for the SM-MM-SM experiment.

a number of 1-m segments as a parameter. Each segment is equipped with rotary splices on both ends. Thus in the one segment configuration, there are two additional splices on the multimode fiber, besides the basic configuration in Fig. 2. The experiment was performed with approximately 2 km of multimode fiber. As the number of splices increases, there is an increase in the BER. However, this increase is relatively small for 1 or 3 segments. Thus our conclusion is that for small number of splices on the multimode filter (say, up to four), the degradation is rather limited.

To evaluate the performance of the scheme as a function of time, the experiment was left running for a period of two weeks. No evident changes in the performance were noticed. However, the scheme may suffer from mechanical perturbations. In other words, the launching and the receiving splices need to be optimized for best performance. Changes in this setting may considerably affect the scheme performance. Moreover, twisting and bending of the fiber may also affect the

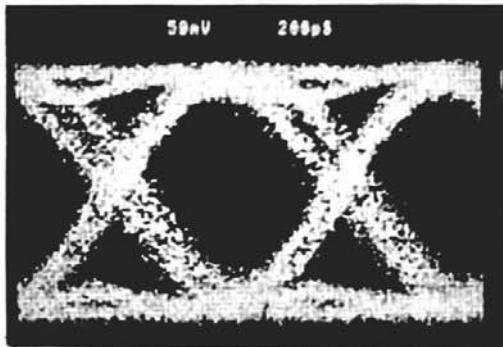


Fig. 19. Eye diagram for the SM-SM-SM experiment.

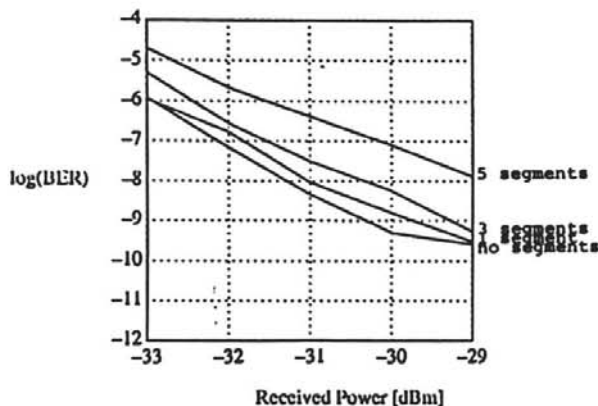


Fig. 20. The effect of splices on the BER performance.

results. Thus it is important that the multimode fiber remains relatively stationary. This would be achieved in wall wiring or in reinforced cables, for example, but may not hold true for loose fiber.

#### IV. SUMMARY, DISCUSSION, AND CONCLUDING REMARKS

The presented scheme extends the *bandwidth-distance* product of a multimode fiber to facilitate upgrading of the existing networks, such as FDDI, wired with multimode fiber, from hundreds of megabits per second to gigabit per second transmission rates. (A similar technique used in a different environment was proposed in [10].) In particular, we have demonstrated that using this technique, 1-Gb/s signal at -30 dBm can be transmitted over 2.3 km of multimode fiber with an error rate of  $1 \times 10^{-9}$ . The penalty of the scheme is approximately 6.5 dB and an improvement of doubling the *bandwidth-distance* product can be achieved, even with multimode fiber with initially large *bandwidth-distance* product. The effect of up to four splices on a 2-km multimode link is relatively limited.

We have concentrated in this paper on the propagation through lower order modes. However, an alternative scheme is possible, in which higher order modes are used. Such a scheme may have, on one hand, the advantage of lower coupling between the modes and, on the other hand, the disadvantage of the more sophisticated launching method. Moreover, because of the lower coupling between the higher order and the lower order modes, and among the higher order modes, filtering at the receiving end may be unnecessary. Since the receiving

filtering is mostly responsible for mechanical susceptibility, the use of the higher order modes for propagation may prove to be a more practical scheme. Also, if no receiving filtering is performed, the power penalty is significantly reduced.

Use of the proposed schemes may have significant economical consequences, when connectivity approaching gigahertz per second rates is needed in the already existing, prewired environments.

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**Zygmunt Haas** received the B.Sc. degree in electrical engineering, in 1979 and the M.S. degree in electrical engineering, in 1985, both with "Summa Cum Laude." From 1979 to 1985 he worked for the Government of Israel. In 1988, he earned the Ph.D. degree from Stanford University, researching fast optical packet-switched networks.

Subsequently, he joined AT&T Bell Laboratories in Holmdel, NJ, where is now a Member of Technical Staff. He is an author of numerous technical papers and holds several patents in the field of optical switching and optical networking, high-speed networking, and wireless networks. He has organized several Workshops and served as an editor for Computer Networks and ISDN Systems journal and as a guest editor for JSAC issues: optical switching, optical switching, mobile communication, and personal communication services.

**Mario A. Santoro** (S'85-M'87) received the professional degree in electrical engineering from the University of Rosario, Santa Fe, Argentina, in 1976, and the M.S., M.Phil., and Ph.D. degrees from the University of Columbia, New York, NY, in 1982, 1985, and 1986, respectively.

In 1976 he joined ACINDAR S.A., Villa Constitución, Argentina, as a Design and Development Engineer where he was involved in the design of control systems. From 1978 to 1980 he worked as an assistant professor in the department of Mathematics, University of Rosario, Santa Fe, Argentina. In 1984, he joined Neteck, Inc., as a research engineer working on lightwave components and systems. In 1987, he joined AT&T Bell Laboratories, Holmdel, NJ, where he conducted research in lightwave networks and systems, and photonic switching applications using WDM. Recently, he joined the newly formed Networking Center of Expertise within AT&T Bell Laboratories, Holmdel, NJ, as a principal architect to provide consultancy services for complete Networking Solutions.

Dr. Santoro is the recipient of the Edwin Howard Armstrong Memorial

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Award for Outstanding Graduate Research. He has authored and co-authored over 30 papers in Lightwave Systems and hold patents on an Optical Packet Switch and on Interleave Receivers. He is a member of SPIE, the International Society of Optical Engineers, the Optical Society of America, and the LEOS representative to PACE, the professional Activity Council for Engineers.

## INFORMATION FOR AUTHORS

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