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ANALOG DATA TRANSMISSION VIA FIBER OPTICS *

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ABSTRACT

In the SLAC Linear Collider Detector (SLD), as in most high-energy particle detectors, the electromagnetic noise environment is the limiting factor in electronic readout performance. Front-end electronics are particularly susceptible to electromagnetic interference (EMI), and great care has been taken to minimize its effects. The transfer of preprocessed analog signals from the detector environs, to the remote digital processing electronics, by conventional means (via metal conductors), may ultimately limit the performance of the system. Because it is highly impervious to EMI and ground loops, a fiber-optic medium has been chosen for the transmission of these signals. This paper describes several fiber-optic transmission schemes which satisfy the requirements of the SLD analog data transmission.

INTRODUCTION

The SLD systems require approximately 170,000 channels of readout electronics.[1] To reduce the cable plant to a reasonable size, it is necessary to multiplex the front-end electronics. For systems such as the Drift Chamber and the Liquid Argon Calorimeter, the data from up to 256 analog front-end channels will be multiplexed onto a single fiber-optic link. The transmission data rate is .67 MHz. As many as 32,000 pieces of analog data will be transmitted per beam period, on a single link, thereby amortizing its cost over many channels. The required dynamic range of the transmission system is 12 bits (4096 : 1).

Three analog data transmission schemes, based upon the Hewlett-Packard HFBR series of fiber-optic components, have been developed for use at SLD. The HFBR series consists of low-cost, high-performance devices which satisfy the system requirements for noise, dynamic range, and bandwidth. Each of these schemes has been optimized for a particular parameter.

The first, optimized for maximum dynamic range, is based upon a discrete transresistance preamplifier. This circuit has demonstrated a dynamic range of >14 bits for $S/N = 1$ and a full-scale output of 10 V. Linearity, however, is approximately ± 10% of reading over the operating range.

The second fiber-optic link described has been optimized for best linearity. This circuit utilizes an integrated circuit preamplifier with optical feedback for linearity correction and stability improvement. The dynamic range has been measured at >12 bits, with a full-scale output of 5 V. Measured linearity is ± 0.5% of reading over the operating range.

A third circuit has been optimized for cost/performance, and takes advantage of the low-cost HFBR plastic series of hybrid PIN diode/preamplifier package. The dynamic range is >12 bits, with a maximum full-scale output of 5 V. Linearity is ± 1% of reading. Because the cost/performance quotient is lower and performance adequate for most applications, this circuit is favored for use at SLD.

OPTICAL LINK DETAILS

Wide Dynamic Range Circuit:

Figure 1 shows a simplified circuit diagram of the optical link transmitter. This circuit, which consists of a voltage to current amplifier and a HFBR–1203 optical transmitter, can couple 1.82 μW/mA typical into a 100/140 μM fiber-optic cable. The HFBR–1203 coupled output optical power (P_T) is related to the input voltage by:

$$P_T(\mu W) = \frac{1.82 \times 10^3 R_{f1} R_{T1} V_i}{R_{S1} R_E (R_{f1} + R_{T1})} \quad . \tag{1}$$

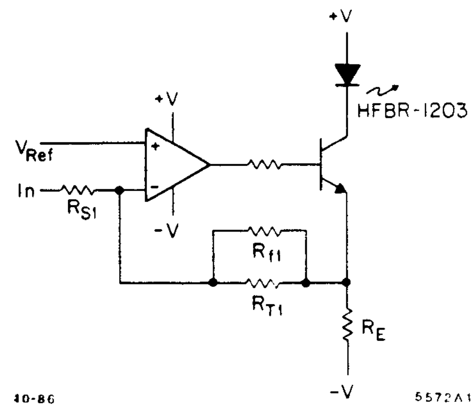

Fig. 1. Transmitter circuit.

In this application, $P_{T\max}$ is equal to 91 μW for $V_{i\max}$. The minimum coupled power is 2.8 nW for a dynamic range of $> 32 \times 10^3 : 1$. R_T is a precision temperature sensor (Micro Switch TD2A) which compensates the transmitter's temperature coefficient.

The fiber-optic loss, 0.17 db for 30 meters plus 1.5 db connector loss at the receiver, results in a received optical power (P_R) of 0.68 P_T.

The receiver circuit (Fig. 2) utilizes an HFBR–2207 PIN diode. When reverse biased, the diode can be represented by the equivalent circuit shown in Fig. 3. The diode noise current (i_N),[2] 9.8 fA/Hz$^{1/2}$ for the HFBR–2207, can be calculated from:

$$i_N \left(\text{A/Hz}^{1/2}\right) = [2qI_D]^{1/2} \quad , \tag{2}$$

where q = Electron charge = 1.6×10^{-19} Coulombs, I_D = Diode dark current = 300×10^{-12}A. R_p for this diode is about $10^{11}\Omega$, which is much greater than R_f, therefore, its noise current contribution can be ignored. A measure of diode noise performance is noise equivalent power[2] (NEP) which is defined by:

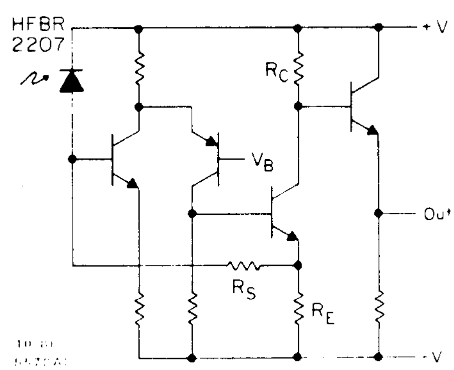

Fig. 2. Wide dynamic range receiver circuit.

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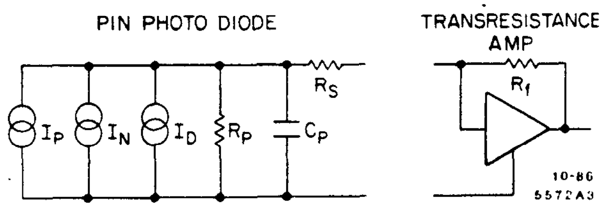

Fig. 3. PIN photo diode equivalent circuit.

$$NEP\left(\mathrm{W/Hz^{1/2}}\right) = \frac{i_N\left(\mathrm{A/Hz^{1/2}}\right)}{R_P(\mathrm{A/W})} \quad , \tag{3}$$

where R_P = diode port responsitivity in amperes/watt. In this application, the minimum required bandwidth is 2 MHz which results in the NEP equal to 36 pW. This is very small compared to the minimum received signal of 0.68 $P_{T\min}$, which is equal to 1.9 nW. It is concluded, therefore, that diode noise is not a limiting factor in this design. Resistor R_f, to which the diode is connected, generates a thermal noise (i_T) equal to 0.2 nA and is given by:[3]

$$i_T = \left[\frac{4KT\triangle f}{R_f}\right]^{1/2} \quad , \tag{4}$$

where K = Boltzman's constant = 1.38×10^{-23} Joule/ °K, T = Room temp = 273°K, $\triangle f$ = System bandwidth = 2 MHz. This is small compared to $0.68 P_{T\min} \times R_P$, which is approximately O.7 nA, and will not contribute significantly to the system noise. In this circuit, the amplifier input transistor determines the dynamic range of the system. The measured noise is 200 μV RMS, which results in a dynamic range of 50×10^3. Specifications for the system are given in Table 1.

Table 1. Wide Dynamic Range Optical Link

SPECIFICATIONS:

ANALOG SIGNAL RANGE = 0 - 10 V OUT

$A_v = -1$

T_r = 225 nS

T_f = 400 nS

NOISE = 200 μV RMS

DYNAMIC RANGE = 50,000 : 1 FOR S/N = 1

LINEARITY = ±10 % of READING FROM 10 mV to 10 V

COST = $250/CHANNEL (1986)

High Linearity Circuit:

The transmitter circuit is identical to the circuit shown in Fig. 1. *Vref*, in this case, is such that 5 mA of quiescent current flows through the optical transmitter diode.

An optical receiver circuit utilizing optical feedback[4] for improved linearity is shown in Fig. 4. The amplifier is a wideband FET input op amp (AD3554), which features a gain-bandwidth product of 1.7 GHz, and is capable of driving ± 100 mA directly.

A simple model which describes the linearizing action of the circuit is shown in Fig. 5a. By superposition, the transmitter coupled power output (P_T) can be represented by the sum of a linear response and a nonlinear response. The nonlinear responses are introduced as separate distortion inputs (P_{D1} and P_{D2}) at summing junctions 1 and 3. L_1 and L_2 represent the transfer ratios of the cables, considering attenuation and the loss at the interfaces to the optical receivers. To simplify the block diagram, $K_1 = P_{T1}L_1R_{P1}R_1$ and $K_0 = P_{T2}L_2R_{P2}R_2$. P_{D1} and P_{D2} are multiplied by K_1/P_{T1} and K_2/P_{T2}, respectively (Fig. 5b). R_1 and R_2 are adjusted, such that $K_0 = K_1$, to compensate for the differences in the current transfer of the two links. P_{D1} and P_{D2} have opposite signs and will cancel for matched transmitter diodes. The diagram may be further reduced to that of Fig. 5c. The ratio of output current to input current is:

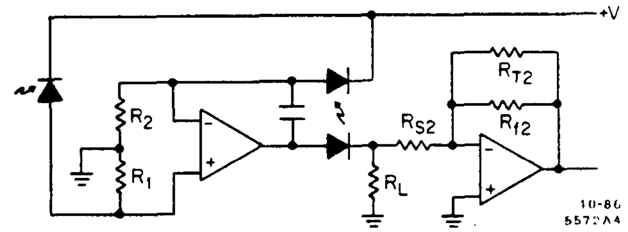

Fig. 4. High linearity receiver circuit.

$$\frac{i_o}{i_i} = \frac{K_1 A}{1 + K_0 A} \simeq 1 \quad . \tag{5}$$

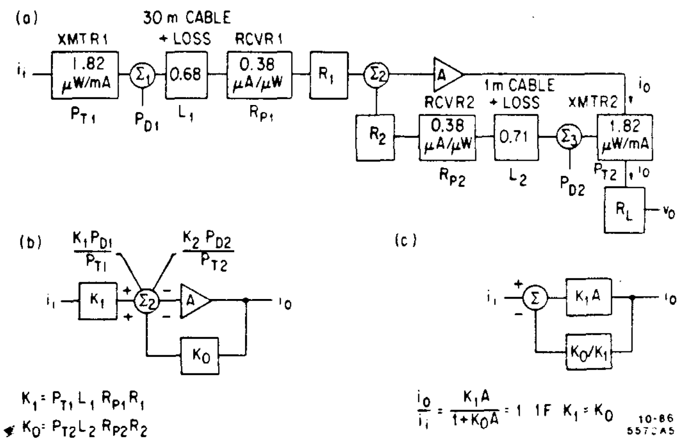

Fig. 5. High linearity receiver block diagram.

if A is large and $K_1 \simeq K_2$. The output voltage is scaled by R_L and the resulting overall voltage gain of the system is:

$$A_v \simeq \frac{R_{f1}R_{T1}R_L}{R_{S1}R_E(R_{f1}+R_{T1})} \times \frac{R_{T2}R_{f2}}{R_{S2}(R_{T2}+R_{f2})} \quad . \tag{6}$$

Specifications for the system are given in Table 2.

Low-Cost Circuit:

The transmitter circuit (Fig. 6) has some minor differences from that previously described. The transmitter diode is a HFBR-1404, which is housed in a low-cost plastic package. This high-efficiency device is capable of launching 195 μW maximum into a 100/140 μM cable and has a transfer ratio of 3.25 μW/mA typical. In this application, the range of coupled power (P_T) is from 16 μW to 59 μW. Manufacturing variations in P_T are as high as ± 150% of the specified

Table 2. High Linearity Optical Link

SPECIFICATIONS:

ANALOG SIGNAL RANGE = 0 - 5 V OUT

$A_v = 2$

t_R = 275 nS

t_F = 275 nS

NOISE = 800 μV RMS

DYNAMIC RANGE = 6250 : 1 FOR S/N = 1

LINEARITY = $\pm$0.5% FROM 10 mV TO 5 V

TEMPERATURE STABILITY = 0.1%/°C + 10°C TO + 50°C

COST = $400/CHANNEL (1986)

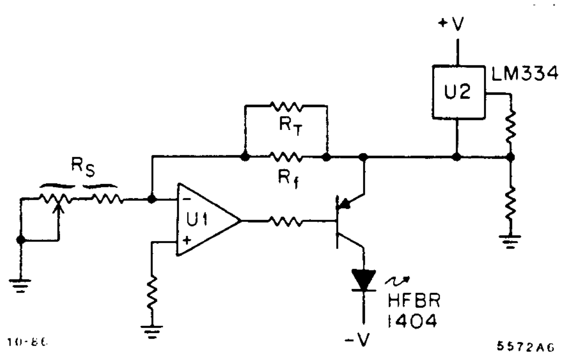

Fig. 6. Low cost transmitter circuit.

typical value. Factory screening of this parameter is available at an increase in cost of 25%, and reduces the spread to ± 20%. This variation in P_T is primarily due to the alignment of the integral optics and connector coupling repeatability. A gain control is necessary to compensate for these variations. The current source, U2, provides 5 mA of quiescent current to the transmitter diode and is independent of the gain control.

The receiver (Fig. 7), a HFBR-2404, contains a discrete PIN photodiode and integrated low-noise preamplifier. This device is also housed in a low-cost plastic package. The responsitivity (R_P) of the HFBR-2404 is 7 mV/μW typical. Factory screening of this parameter is also necessary. The minimum detectable signal in the system is determined by noise, which is a function of the system bandwidth. The minimum system bandwidth is determined by a readout rate and the step response settling time to a given accuracy. Assuming a non-return to zero (NRTZ) data rate of 1.5 μS per analog sample and a settling time of 1.3 uS to 0.1%, the minimum system bandwidth is approximately .8 MHz. This is determined by:

$$\tau = \frac{-t}{\ln\left(1 - \frac{v}{V}\right)} \quad , \tag{7}$$

where τ = system time constant in sec, t = settling time to desired accuracy, v = instantaneous voltage at t in volts, V = final voltage in volts, and

$$BW_{\min} \simeq \frac{.35}{t_r} = \frac{.35}{\tau\left[\ln\left(1 - \frac{v_1}{V}\right) - \ln\left(1 - \frac{v_2}{V}\right)\right]} \quad , \tag{8}$$

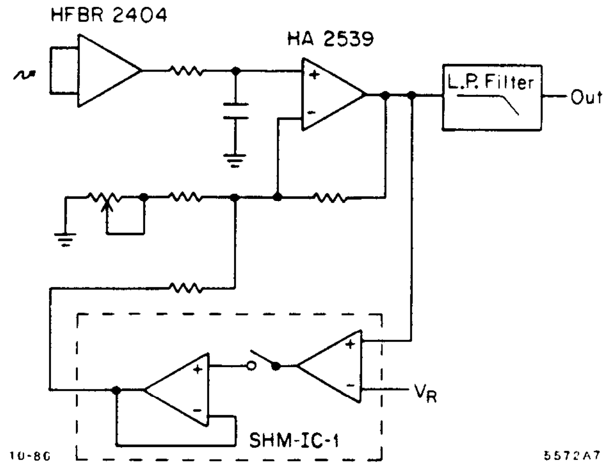

Fig. 7 Low cost receiver circuit.

where $BW_{\min}$ = minimum system bandwidth in Hz, t_r = rise time (10–90 %) in sec, V_1 = 0.1 V, V_2 = 0.9 V. Combining and simplifying

$$BW_{\min} \simeq \frac{-.35 \ln\left(A \times 10^{-2}\right)}{2.2\ t} \quad , \tag{9}$$

where A = settling accuracy in %. The noise voltage out (E_N) for the HFBR-2404 is 72 nV/Hz$^{1/2}$ maximum at 25°C and results in an output noise of 0.1 mV, for a 2 MHz system bandwidth. The maximum linear output swing is 500 mV, which allows a dynamic range of 5000 : 1. This sets the upper bandwidth limit to 2 MHz, while the lower limit is .8 MHz. To achieve the required noise and settling time performance, the bandwidth is limited to 1.4 MHz by a low-pass filter, at the postamplifier output. The postamplifier is a HA-2539, a low-cost, fast-settling (10 V step to 0.1 % in 200 ns) operational amplifier. Its nominal noninverting gain, in the signal path, is 11.

The DC stability of the HFBR-2404 integrated preamplifier, several mV/°C, would not normally be adequate for this application. An adaptive auto-zero circuit, to improve DC drift performance, consists of a SHM-IC-1, sample-and-hold, in a postamplifier feedback loop. During system deadtime, the sample-and-hold is put into sample mode. The output of the postamplifier is compared to the reference voltage at the negative input of the sample-and-hold. This difference is driven to zero, due to the high, open-loop gain of the sample-and-hold input amplifier, regardless of the DC offset voltage at the output of the HFBR-2404. A few microseconds before data is to be transmitted, the sample-and-hold is put into hold mode. The DC correction voltage stored on the sample capacitor maintains the correction during the readout period. The readout period is limited by the sample-and-hold droop.

Periodic calibration of the system gain will be facilitated by the use of a calibrate signal injected into the detector preamplifiers. This is necessary to correct for errors introduced by the optical transmitter and fiber degradation.

Specifications for the system are given in Table 3.

Table 3. Low Cost Analog Optical Link

SPECIFICATIONS:
ANALOG SIGNAL RANGE = 0 - 5 V OUT
$A_V = -2$
$t_R = 250$ nS
$t_f = 250$ nS
NOISE = 800 μV RMS
DYNAMIC RANGE = 6250 : 1 FOR S/N = 1
LINEARITY = ±1 FROM 10 mV TO 5 V
TEMPERATURE STABILITY = 0.1%/°C + 10°C TO + 50°C
COST = $175/CHANNEL (1986)

CONCLUSION

Three analog optical links have been designed and tested at SLAC. The dynamic range of the links, up to 50,000 : 1, is not limited by the optical component performance. It is ultimately limited, however, by the optical receiver preamplifier noise. Linearity can be improved, at the expense of dynamic range, by the use of optical feedback in the transmission system. The low-cost link has been chosen for use at the SLD. This is mainly due to the reduced dynamic range requirement brought about by the LAC dual-ranging scheme.[1]

New devices, such as the HFBR series of low-cost optical devices, are finding use in modest cost versus performance systems. As the fiber-optic technology matures, higher performance and cost-effective devices will make analog data transmission a more attractive alternative to transmission over metallic conductors.

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