FIBER OPTIC TRANSMISSION

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Fiber optic transmission is assuming an increasingly important role in systems for wide-band analog signals and digital signals with high data rates. Although the number of applications for digital networks and telecommunications systems is skyrocketing, analog transmission is still vital to many applications. Analog systems with bandwidths of up to 150MHz are used for wide-band RGB signal distribution, HDTV video signal transmission, and many types of EMI- and EMC-disturbed environments. Also important are medical applications, which demand the precision of fiber optic technology for safety reasons. The many features of fiber optic cables make them vital for all of these types of applications. Fiber optic cables enable transmission over long distances, ensure low damping vs frequency, are light and flexible, and provide high immunity against disturbances from magnetic and electric fields. State-of-the-art fiber optic transmission systems are now available even for data networks with transmission rates of up to 1.2Gbit/s, and gallium arsenide technology is used for their transmitter and receiver circuits.

The fiber optic transmission interface presented here uses new complementary bipolar integrated circuits from Burr-Brown. The OPA660, which is used as an LED driver and AGC multiplier, contains an operational transconductance amplifier and a buffer in an 8-pin package. The OPA621 is a low-noise, wide-band op amp in classical configuration, which functions as an amplifier in the I/V conversion section behind the photodiode and as an I/V converter behind the AGC multiplier. The current-feedback amplifier OPA623 provides additional gain in the AGC section and drives the 75Ω output. A discrete differential amplifier functions as an AGC error amp and controls the quiescent current of the OPA660 together with a FET. The CA3080 stabilizes the DC performance, and the LM1881 functions as a sync separator. The interface uses the IA184A as LED and the SFH202 as pin diode.

FIBER OPTIC INTERFACE BASICS

A fiber optic interface generally consists of five major functions as shown in Figure 1. On the transmitter side, a circuit processes the input signal in order to drive the electro-optical converter. This converter, which can be an LED or a laser diode, generates the signal-dependent light intensity modulation, and its mechanical case eases transmission of the signal into the fiber. At the fiber end, a pin diode converts the optical signal back into a low electrical current. The low-noise transimpedance preamplifier converts the current signal into a voltage and also amplifies it to an acceptable level.

Because the photodiode input signal can vary in amplitude, and AGC amplifier adjusts the peak-to-peak signal level to 1.4Vp-p and restores the DC level for no signal to 0V.

The quality of a fiber optic interface is characterized by several factors such as signal-to-noise ratio, linearity, bandwidth, power consumption, and transmission distance. The S/N ratio should be at least 50dB for analog systems to achieve an image that is free of noise. In conventional designs, there are basically two ways to improve the S/N ratio. One is to increase the diode drive current, which, though, leads to higher harmonic distortion. The second is to use a very low-noise transimpedance amplifier as a receiver. Both alternatives increase the component count and add manufacturing costs. The circuits presented here, however, are a new approach to simplifying and minimizing the design of an analog fiber optic interface and to provide an interface that is more integrated and offers lower power consumption. Table I summarizes the parameters of this new type of wide-band analog fiber optic interface.
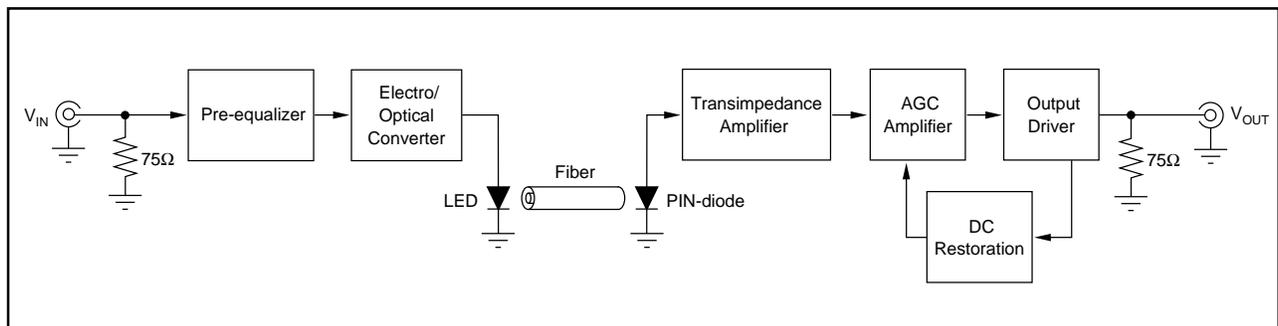
FIGURE 1. Block Diagram of a Fiber Optic Transmission Interface.

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PARAMETER	UNIT
Bandwidth	120MHz
Differential Gain	≤ 3%
Differential Phase	≤ 3°
S/N Ratio	≥ 50dB
AGC Range	20:1
Input Voltage	+0.7V/–0.3V terminated in 75Ω
Output Voltage	+0.7V/–0.3V terminated in 75Ω
Supply Voltage	±5V

TABLE I. Interface Parameters.

TRANSMITTER

The block diagram illustrated in Figure 2 can be divided into three major blocks. The preequalizer compensates for the nonlinearity of the diode. The driver circuit converts the input signal into an output current, which generates the optical signal when flowing through the LED. For the LED to function linearly in the forward region, a positive DC current has to flow through it to adjust its bias point and keep it constant over temperature variations. Figure 3 shows the discrete circuit to adjust the LED bias point. R_{QC} can be calculated by the following equation:

$$R_{QC} = \frac{V_+ - V_{BEQ2}}{I_{EI}} - R_{D2} \qquad (1)$$
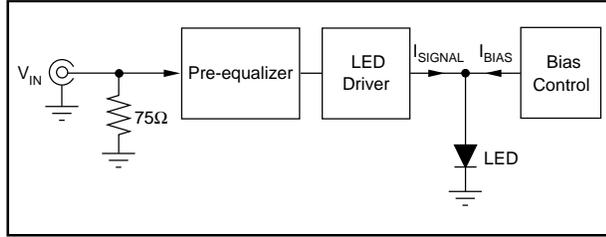
FIGURE 2. Block Diagram of Transmitter.
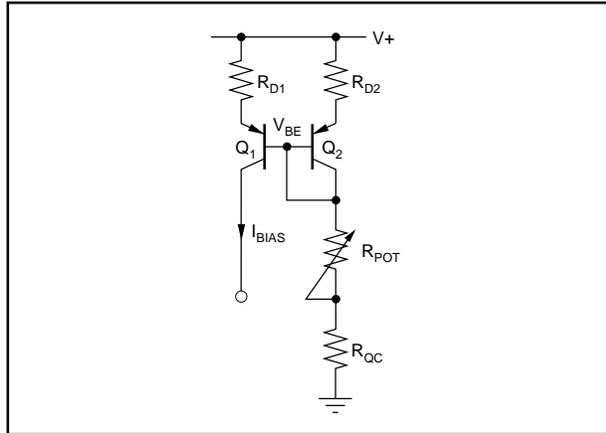
FIGURE 3. Bias Control.

One characteristic of a current source is its high impedance. To avoid current distribution, the output impedance should be much higher than the load impedance. The transmitter diode is a 2Ω to 4Ω load adjusted to the correct bias point.

The high output impedance of 10kΩ for the current source prevents any distribution.

To modulate the diode current, the OTA of the OPA660 operates as a voltage-controlled current source and converts the input voltage into output current. As shown in Figure 4, the resistor R_E is the only element that has to be selected in order to define the conversion factor between input voltage and output current according to equation 2:

$$R_E = \frac{V_{IN}}{I_{OUT}} - R_e \qquad (2)$$
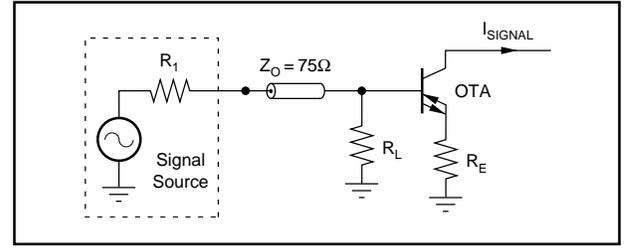
FIGURE 4. Voltage-to-Current Converter.

R_E is the output impedance of the OTA's emitter output and is 8Ω or 1/125mA/V at a +20mA quiescent current. The transfer curve between the diode current and optical power is only fairly linear within a small modulation range around the bias point. The larger the modulation range, the larger the nonlinearity. The equalization circuit proposed in Figure 5, however, can be used to improve the linearity or extend the modulation range. The diodes function like switches and connect the resistors in parallel to R_E when the corresponding diode is forward biased. The equalization circuit varies the voltage-to-current conversion according to the input voltage factor (mV/A) and partly compensates the diode nonlinearity.
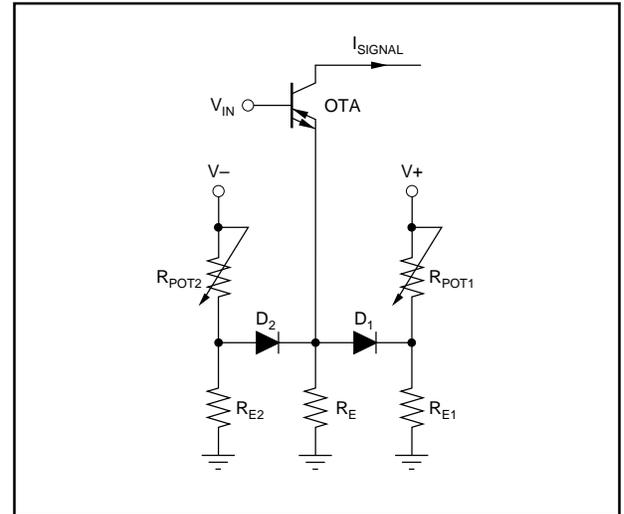
FIGURE 5. Transmitter with Preequalization.

As shown in Figure 6, the differential gain of the entire transistor circuit decreases from 25% without equalization to 6% with equalization. With more hardware, a further improvement down to 2% would be feasible. Figure 6 also presents the differential gain errors for the diode current, which impressively demonstrate that the conversion from a current to an optical power generates most of the nonlinearities.
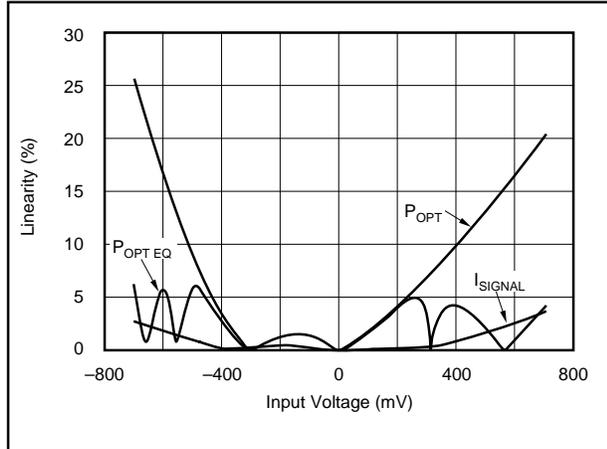
FIGURE 6. Transmitter Linearity Performance.

The complete transmitter is shown in Figure 7. Using the component values given in Figure 7, the discrete biasing circuit adjusts the diode current without modulation to +35mA, which produces a +1.55V voltage drop across the diode. Even at the highest modulation current of 30mA, the voltage drop of +1.7V remains far below the collector-emitter saturation voltage.

The drive capability of the OPA660 is limited to ±15mA. For this reason, two OTA current source outputs are connected together to increase the drive capability to ±30mA.

The input signal is applied to both bases of the OTA. In this application, each OPA660 operates with 20mA quiescent current, and requiring a R_{QC} resistor of 250Ω. The unused buffers of the OPA660s can either be connected to GND by a resistor or used for other circuit functions such as compensation of the OTA input offset voltage (in this case, R_E should be connected to the buffer output).

Figure 8 shows the frequency response of the diode current and optical power. Equation 3 gives a reasonable calculation for the 3dB frequency based on the effective capacitance at the diode anode and the total resistance:

$$f = \frac{1}{2\pi R\left(C_P + C_{DT}\right)} \qquad (3)$$

R = Diode resistance, 3Ω
C_P = Biasing circuit, 25pF
C_{DT} = Input capacitance OTA-C, 8pF
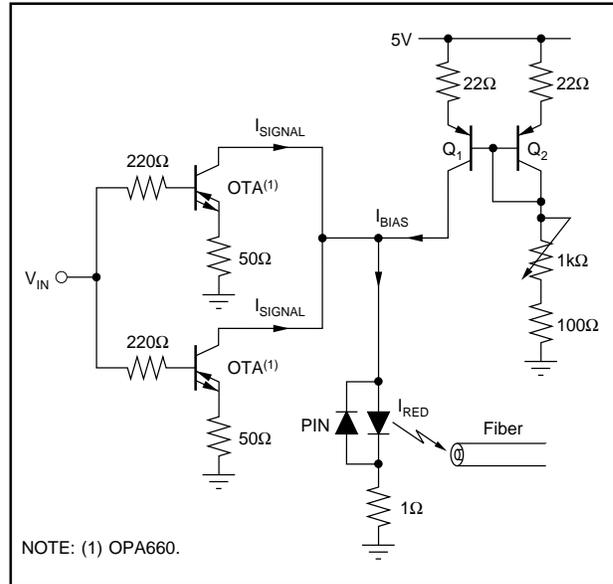
FIGURE 7. Transmitter Circuit.
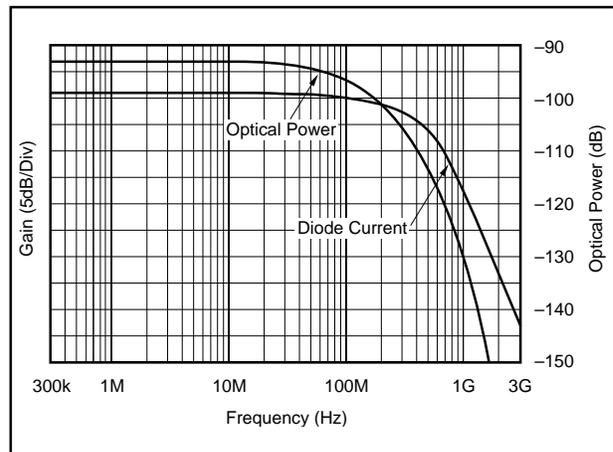
FIGURE 8. Bandwidth Transmitter.

When these values are put into the equation, the transmitter has a –3dB bandwidth of 800MHz, which corresponds to that in actual measurements. Comparing the curves in Figure 8 with each other shows that the transmitter diode and not the transmitter is the bandwidth-limiting factor. However, the 115MHz optical power matches the figures provided in the diode specification.

LOW-NOISE TRANSIMPEDANCE AMPLIFIER

On the receiver side, the electronic circuitry converts the optical power into a voltage, amplifies the normally weak signal, and stabilizes the output voltage for different cable lengths via an AGC control loop. A sensitive, small, fairly linear PIN-diode such as the SFA202 used here delivers output currents in the μA to mA range in typical applications. To convert weak, wide-band signals into voltages, while simultaneously amplifying them, is a tough job for

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wide-band amplifiers. Figure 9 shows the typical transimpedance configuration. The PIN-diode cathode is directly connected to the inverting op amp input. The positive input is tied to GND. The effective transimpedance resistor connects the output to the inverting input. For low frequencies, the output voltage is calculated by:

$$V_{OUT} = -I_P \bullet \frac{R_F}{1 + 1/G_{OL}} \approx -I_P \bullet R_F \qquad (4)$$
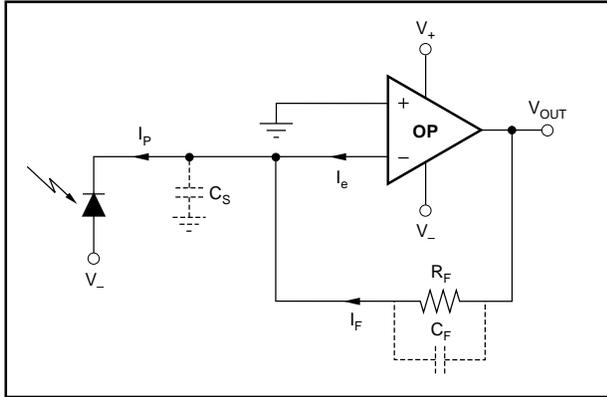
FIGURE 9. Transimpedance Amplifier.

Obviously, amplification does not work at higher frequencies where the open-loop gain (G_{OL}) decreases. Op amps, like the OPA621 used here, are internally compensated, and the open-loop gain rolls off by –20dB/decade.

Equation 5 allows an estimation of the maximum R_F size for a given –3dB bandwidth.

$$R_F = \frac{1}{2\pi 100MHz \bullet 0.5pF} \approx 3.2k\Omega \qquad (5)$$

The minimum detectable input current depends upon the noise performance, which in turn is based upon the noise of the low-noise preamplifier. The various noise sources of a transimpedance amplifier are shown in Figure 10. After the op amp noise, the next most important noise factor is the total thermal noise from all of the resistances. The thermal noise can be calculated by the following equation:

$$V_{th}^2 = 4 \bullet kT \bullet R \bullet B \qquad (6)$$

Thus, the effective noise voltage increases with temperature, T, bandwidth, B, and resistor size. Another often negligible noise factor is the pin diode itself. The diode's noise is mostly 1/f noise and is much higher in the forward than in the reverse region of the diode biasing. It can be calculated as follows:

$$i_s^2 = 2eI_P \bullet B \qquad (7)$$

To analyze the noise from all effective noise sources at the transimpedance amplifier output, determine the effective noise from all of the individual sources and add them
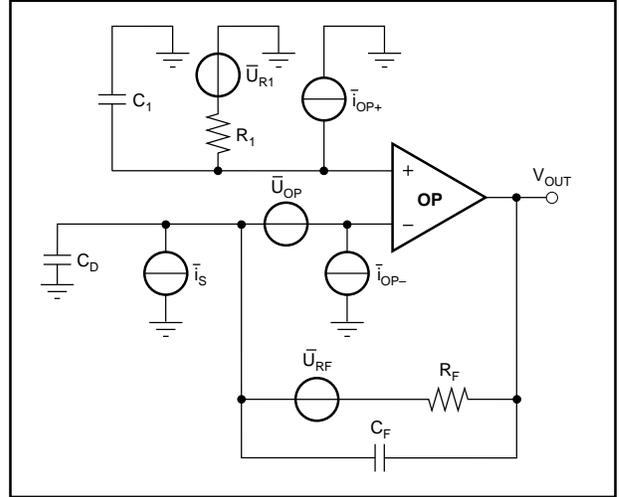
FIGURE 10. Noise Sources of a Transimpedance Amplifier.

geometrically to find the square sum. The purpose of a noise analysis is to ascertain the minimum detectable input signal current from the pin diode. The preamplifier noise performance also determines the required transmitter power, maximum cable length for a given transmitter current, dynamic range of the receiver, and signal-to-noise ratio. By dividing the voltage noise at the output by the transimpedance resistor R_F, it is possible to calculate the equivalent input noise current, which is also the minimum detectable input signal.

The noise performance of any op amp varies over frequency, because in any integrated circuit, different noise sources are effective in different frequency ranges.

Table II summarizes the voltage noise density and effective noise voltage for the OPA621 vs frequency. The input current noise density is listed as 3.3pA/$\sqrt{Hz}$ in the PDS. Table III adds up the single noise sources to find the effective noise voltage at the amplifier output for three different frequencies. It can be derived that the equivalent noise voltage is dependent upon the system bandwidth. Thus, for any given transimpedance amplifier, the S/N ratio decreases with increasing bandwidth.

FREQUENCY/Hz	VOLTAGE NOISE/nV/$\sqrt{Hz}$	EFF. NOISE/μVrms
0 - 200	10	0, 141
200 - 2k	5, 5	0, 23
2k - 20k	3, 3	0, 44
20k - 1M	2, 5	2, 47
1M - 10M	2, 3	6, 90
10M - 100M	2, 3	21, 82
TOTAL		24.17

TABLE II. OPA621 Noise Performance vs Frequency.

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NOISE SOURCE	VOLTAGE NOISE DENSITY nV/√Hz	EFFECTIVE VOLTAGE NOISE μVrms		
		5MHz	10MHz	100MHz
$\overline{u_{R_F}^2}$	4, 07	9, 1	12, 9	40, 7
$\overline{u_{R_1}^2}$	1, 29	2, 9	4, 1	12, 9
$\overline{u_{OP}^2}$	2, 42	5, 4	7, 7	24, 2
$\overline{i_{OP+}^2}$	0, 33	0, 7	1, 0	3, 3
$\overline{i_{OP-}^2}$	3, 30	7, 3	10, 4	33, 0
$\overline{i_S^2}$	1, 79	4, 0	5, 7	17, 9
TOTAL		13, 8	19, 6	61, 9

TABLE III. Receiver Noise Performance.

As already stated, dividing the equivalent noise at the output by the transimpedance resistor value produces the equivalent input current noise. The following equation is used to calculate the S/N ratio, and Table IV shows the S/N ratio vs frequency:

$$S/N = 20\log\left(\frac{I_P}{I_{RMS}}\right) \qquad (8)$$

FREQUENCY (MHz)	CURRENT NOISE (nA)	SIGNAL-TO-NOISE RATIO (dB)
5	13, 8	57, 2
10	19, 6	57, 2
100	61, 9	44, 2

TABLE IV. Signal-to-Noise Ratio.

When low-noise, wide-band amplifiers are combined with very low-noise discrete FETs, the results are not only lower noise but also a higher S/N ratio and only slightly smaller bandwidth. As shown in Figures 11 and 12, the discrete J-FET J308 at the input of the final transimpedance amplifier version improves the S/N ratio by about 5dB at 100MHz. The FET in front of the OPA621 functions as a source follower, but it is connected to the OPA621 and feedback loop in such a way that it does not change the basic impedance structure. The photon current flows to the FET gate and generates a voltage change. The source follower transfers the voltage variation to the noninverting input at a lower impedance. In its negative feedback loop, the op amp also reacts by varying its output voltage, which causes a current to flow through the feedback network until the source voltage equals the voltage at the noninverting input.

The source follower provides no voltage gain but enough current gain that the noise from the OPA621 is negligible. Since the first gain stage in a composite amp is the primary noise-producing element, the main noise source of the circuit presented here is the input current noise at the FET gate, which is about 1pA/√Hz for the J308 at 100kHz.
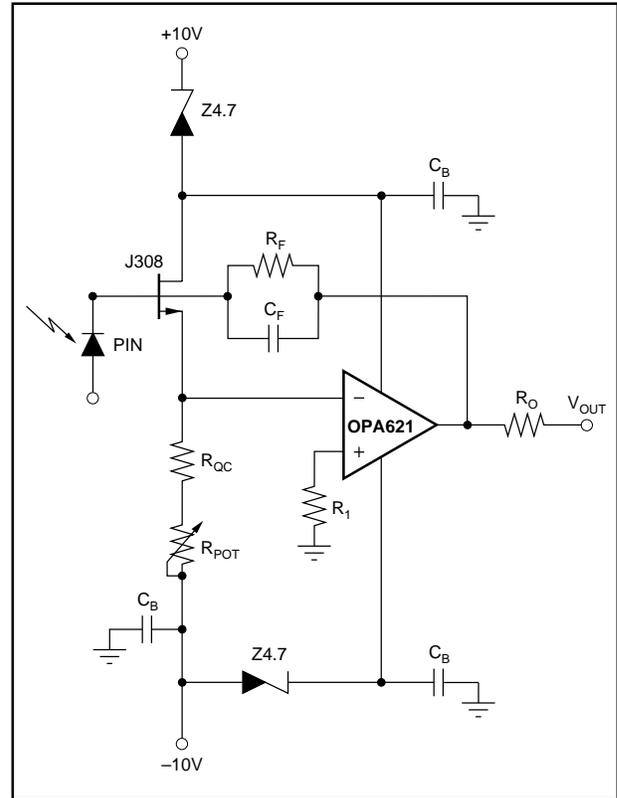
FIGURE 11. Receiver Circuit.
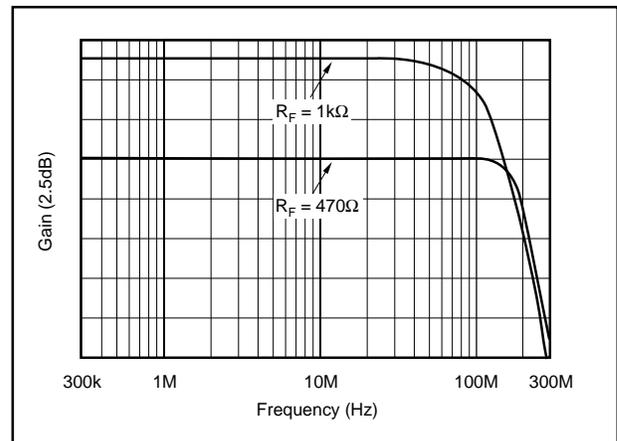
FIGURE 12. Receiver Bandwidth.

PARAMETER	VALUE	UNITS
Frequency	100	MHz
V_{NOISE}rms	36	μV
I_{NOISE}rms	36	nA
R_F	1	kΩ
S/N	48.9	dB

TABLE V. Summary of Noise-related Performance at 100MHz.

The resistor R_{QC} enables the user to adjust the FET bias point. To test the frequency response at the PIN-diode cathode, a generator provides a 10µA photo current, and an analyzer records the output vs frequency. The diode anode is connected to –15V in order to minimize the effective stray capacitance, which remains 5pF.

Two frequency response curves are shown in Figure 12. The top curve reflects the measurement using a 1000Ω transimpedance resistor; here the –3dB bandwidth is 110MHz. The bottom curve is for a 470Ω resistor, where the –3dB bandwidth is about 200MHz.

Table VI shows some test results for the fiber optic transmission system when the input signal is applied to the transmitter. The transmitter supplies the LED with a 35mA quiescent current and a ±25mA modulation current.

CONDITIONS	BW (MHz)	S/N (dB)	DG (%)	THD (%)
R_F = 470Ω w/o Preequalization	135	40.9	27	—
R_F = 1kΩ w/o Preequalization	100	44.2	27	—
R_F = 1kΩ With Preequalization	100	44.2	6	1

TABLE VI. Performance Summary.

The noise of the transmission system between 0Hz and 200kHz is illustrated in Figure 13.
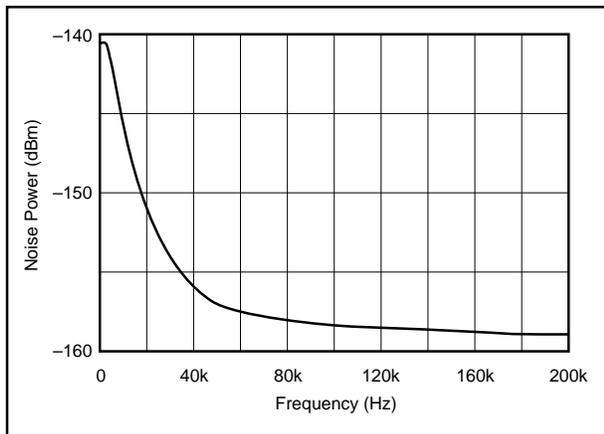
FIGURE 13. Noise Power vs Frequency.

AGC AMPLIFIER AND LINE DRIVER

Multiplication of analog signals has long been one of the most important nonlinear functions of analog circuit technology. Many signal sources, however, deliver weak, oscillating, and simultaneously wide-band signals. The PIN-diode current presented here is no exception, but it is equipped with an AGC amplifier to solve this problem. The AGC amplifier, in this case the OPA660, measures the output voltage, compares it to a reference voltage, and adjusts the multiplier control until the output has reached the set value.

The OPA660 is used in this configuration both as a two quadrant multiplier and as an amplifier. By varying its own gain, it keeps the output constant over a wide input voltage range. Figure 14 shows a simplified circuit diagram of the AGC configuration. The output signal of the transimpedance amplifier, which is 10mV for a 10µA input current and a 1kΩ R_F, is applied to the buffer input of the OPA660. It is configured as a differential amplifier with current output. The second input allows DC restoration of video signals, as will be shown later. The amplifier (OPA621) placed after the OPA660 converts the output current i of the multiplier into a voltage, while providing additional gain and drive capability. The peak detector and comparator compare the typical ±1.4V video output voltage with the +1.4V reference. The resulting difference in voltage controls the gate of the FET. The gate varies the FET drain current, which is also the OPA660 quiescent current, until the internal OPA660 transconductance has compensated the varying input voltage.
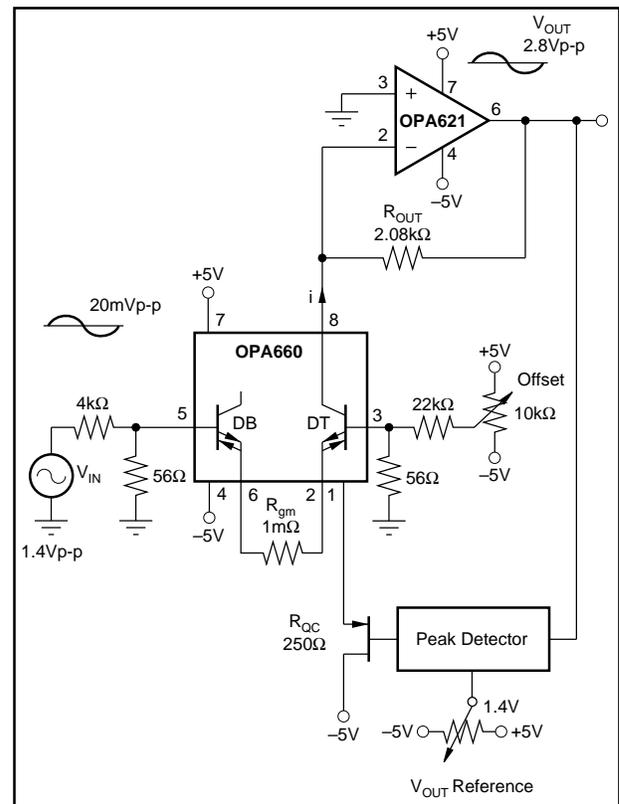
FIGURE 14. Automatic Gain-Controlled Amplifier.

Figure 15 shows the detailed AGC amplifier schematic. The output amplifier is split into an OPA621 and OPA623, reducing the gain requirement for each individual op amp in order to increase the achievable bandwidth. Unfortunately, it is not feasible to replace the voltage-feedback amplifier OPA621 with a wider bandwidth current-feedback op amp because the feedback resistance is 20kΩ. In current-feedback amplifiers, the size of the feedback network determines both the closed-loop and the open-loop gain. Thus in prac-

tice, the larger the resistor, the lower the bandwidth. The peak detector and comparator are made up of discrete components. For a more integrated solution, the differential amplifier can be replaced by an op amp.

Besides the automatic signal control, the circuit presented here also makes it possible to control and adjust the DC level of the output voltage. This feature is useful for video applications in which the blank level, or level at no luminance signal, is defined as 0V. The sync signals are defined from zero to –0.3V, and the luminance ranges from zero to +0.7V. To transmit a video signal over a coax cable, an amplifier amplifies this signal by 2. The whole procedure is called DC restoration or black level clamping. A video signal is a signal that appears periodically between sync pulses. The signal remains at black level for a short time after each horizontal signal pulse, which controls the line information. During this short time, a gated error amplifier compares the output level with the reference voltage (GND) to correct the output to GND. The OPA623 amplifies the signal from the preamp and drives the sync separation circuit LM1881. It provides the necessary clamp pulse shortly after the II sync. The buffer, BUF601, applies the clamp pulse II_C to the restoration circuit CA3080, which is switched on and generates the correction voltage that is then stored in C_4. The comparison stops when the clamp pulse returns to logic "low", but the capacitor keeps the output voltage at GND as a biasing point until the next clamp pulse is applied.
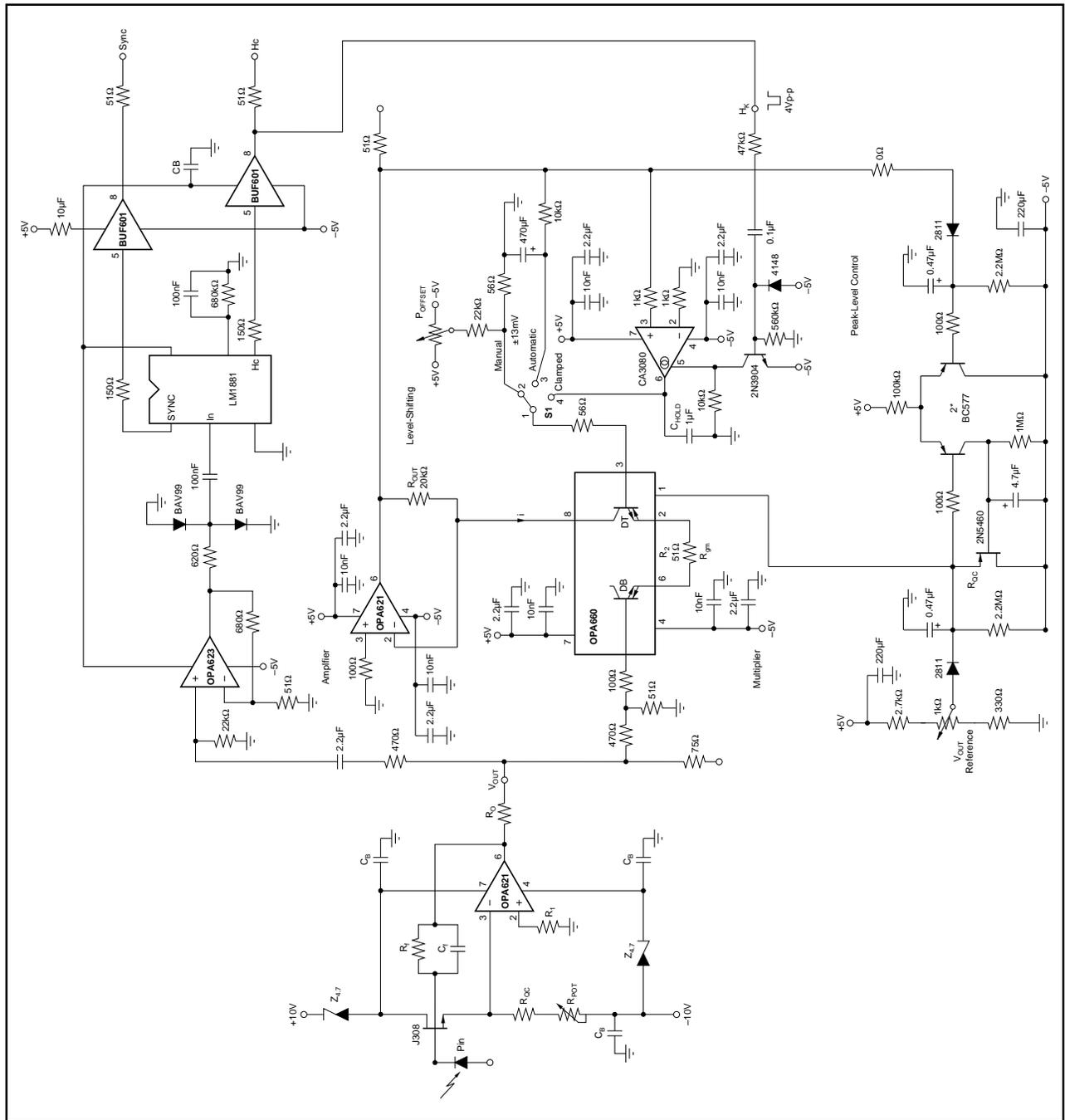
FIGURE 15. Circuit Schematic of the Transimpedance and AGC Amplifiers.

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