Low cost, TTL fiber optic receivers for up to 100Mb/s NRZ AN1443

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INTRODUCTION

This application note describes circuits belonging to the receiver part of the Philips Semiconductors HiFI-100 (High-performance Fiber-optic Interface) Fiber Optic Chip Set. It should be noted that the performance of each board reflects the capabilities of the system along with its external components and does not represent the maximum capabilities of the individual Integrated Circuits. Performance may vary depending on layout and/or external component values.

Because of the advantages offered by fiber optics, such as high bandwidth, EMI immunity, light weight, etc, many users are switching from coaxial or twisted pair links to optical fiber. However, since most systems are not yet able to take full advantage of the data bandwidth of fiber (up to several Gb/s), users are looking for an inexpensive migration path to optical fiber. They would like to make the switch, but they want unit costs to be competitive with those of copper links. They also want enough flexibility so that if they want to increase their data rate, the upgrade costs in hardware would be minimal. The Philips Semiconductors HiFI-100 chip set lets users create inexpensive data links for transmission rates up to 100Mb/s NRZ (typical).

The HiFI-100 series comes in several flavors, each flavor describes a different combination of preamplifier and postamplifier. The most popular combinations are the HiFI-100A, HiFI-100B, and HiFI-100C. These are defined as follows:

Chip-Set	Preamp	Postamp
HiFI-100A	NE5211	NE5214
HiFI-100B	NE5210	NE5217
HiFI-100C	NE5211	NE5217
HiFI-100D	NE5212	NE5217

Each combination has its own advantages and disadvantages with respect to sensitivity, bandwidth and RZ or NRZ operation. Deciding which combination to use is the subject of the section on systems considerations. Experimental results on the HiFI-100C will be presented later.

System Considerations

Figure 1 shows a typical point-to-point fiber optic information channel. Assuming that the data was sent out on a parallel bus (8, 16, or 32-bits wide), it must be converted to a serial stream of data. At this point it may be encoded to optimize transmission. To reduce bandwidth, it may be changed from RZ to NRZ. To facilitate clock recovery it may be converted from NRZ to NRZI or Manchester, which would increase the bandwidth, so there are several trade-offs to consider. After the encoding stage, the signal goes to an electrical-to-optical converter. From there the information (as light) travels over the fiber medium. For lower speed applications, multi-mode fiber cables and LEDs of 850nm wavelengths are used. For higher bandwidth, longer distance applications, single-mode fiber is the cable of choice. For transmitting at these frequencies (up to several Gb/s), lasers and light of 1300nm or 1550nm are used .

On the receiving side, we have the reverse set up. An optical-to-electrical converter is used to raise the signal to either TTL or ECL levels. Also, the signal must be processed to recover the clock which had been embedded during the previous encoding stages. This recovered signal is then sent to the decoding section and then to the next logical section, serial-to-parallel conversion.

A Typical Fiber Optic Data Link

The domain of the HiFI-100 chip set covers the E/O, O/E and the clock recovery sections as shown in the blocks above. Figure 2 shows how this breaks down to functional blocks. In the transmitter, one block takes the serial, encoded signal and uses the signal to drive an LED. In the receiver, a photodiode accepts the light signal and converts it into a current which is then input to the preamplifier. This preamplifier converts the signal into a differential voltage which is then input to a postamplifier which shifts the signal to an appropriate logic level (TTL for the HiFI-100). The clock recovery section recovers the clock from the transitions in this signal, and then re-times the data for further processing.

The focus of this application note will be on the receiver section, excluding clock recovery. We will concentrate on the preamplifier and postamplifiers, and on the trade-offs involved with each combination.

Noise Considerations

The most important noise sources to consider are at the front end of the receiver since successive stages will only amplify whatever noise occurs here. The preamplifier is primarily concerned with two types of noise: that which comes from the preamplifier, itself, and the noise that comes onto the input current from the photodetector. Several publications provide detailed explanations of these effects.

Data Patterns

The type of data pattern input to the receiver will play a large role in determining the performance or sensitivity of the receivers. For the HiFI-100 series, the type of postamplifier chosen depends on the type of data pattern received. The following analysis will provide a guide to choosing the proper postamplifier.

The choice of preamplifier will be determined primarily by user requirements on sensitivity, noise, dynamic range and bandwidth. The tradeoffs associated with each of the three preamplifiers will be discussed in the preamplifier section.

To understand the types of data patterns being transmitted, Figure 3 shows the types of patterns considered. On the top row we have the data that we wish to transmit. Each ‘1’ or ‘0’ occupies a single bit cell. If we assume that a voltage level V+ corresponds to a logic ‘1 ’ and that a level V- corresponds to a logic ‘0’, the total swing is 2V.

Manchester

Manchester encoding is most often recognized by its regular transitions in the middle of each bit cell. A logic ‘1’ is characterized by a low-to-high transition in the middle of each cell while a logic ‘0’ has a high-to-low transition. The advantage of Manchester is that the regular transitions make it ideal for clock recovery systems. One disadvantage is the bandwidth requirement. For a 50MHz system, it can transmit only 50Mb/s maximum. An equivalent NRZ system can transmit twice the number of bits, giving Manchester only 50% of the efficiency of NRZ.

Return-to-Zero (RZ)

The significance of RZ data is that for a logic ‘1’ it stays high for the first half of the bit cell, but returns to the logic ‘0’ level by the end of the bit cell. For logic ‘0’ it stays at the zero level throughout the bit cell. For long strings of zeroes, the input signal resembles a DC-level, which would cause problems in AC coupled systems.

Non-Return-to-Zero (NRZ)

The major difference between NRZ and RZ is that the logic level stays the same throughout the bit cell. This means that the

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maximum transmission rate for the data in bits/sec is just twice the maximum frequency transmitted in Hz. For example, the highest frequency transmitted in a 100Mb/s NRZ system would be a 50MHz square wave. Potential problems with NRZ systems occur with long strings of ones and zeroes. These strings placed back-to-back look like low-frequency signals that could cause problems in AC-coupled systems. Because of the relative lack of transitions, clock recovery would also be more difficult.
NRZI is a variation of NRZ where a “1” is designed by either a low-to-high or high-to-low transition.

Low Frequency Operation

Low-frequency cut-off is another problem that could be considered data-dependent. This usually occurs in RZ or NRZ systems with long strings of ones or zeroes . For RZ, only the long strings of zeroes could cause problems. For NRZ, both logic states could introduce problems. Clearly, the larger the number of bit cells before a transition occurs, the more it looks like a low-frequency signal.
For systems with Auto-Zero Loops (such as the NE5214 and NE5217), these lower frequency signals are rejected as input offset when they should be passed through as valid information. This is not to say that systems with Auto-Zero loops are bad because in cost sensitive applications, saving the cost and board space of an additional coupling capacitor (one Auto-Zero capacitor versus two coupling capacitors) is significant, and the AZ-loop takes out DC offsets that would be amplified.

In general, users should ask themselves what the lowest frequency possible sent will be given the transmitter’s encoding scheme. Based upon this, they can then choose what value of Auto-Zero capacitor to use for sensing the lower-frequency pole or what value of coupling capacitors to use in an AC-coupled system.

AC-coupling

The advantage of AC-coupling when using components from different manufacturers is clear: it lets you use components that have different DC-bias levels. This is often the case when a user wants to mix and match different preamplifiers and postamplifiers because of the differences in their input and output bias levels.

Unfortunately, AC-coupling has its own set of disadvantages. The first would be component cost and space for the two capacitors needed. In hybrid applications where the physical dimensions of the data link is critical, this could pose a major problem. Another problem typically associated with low frequency pseudo-random data is baseline wander.

Looking at Figure 4, we have the same waveforms described above, but with their DC averages superimposed on them. With the transmitted data patterns, as the DC average (or baseline) wanders or moves, the switching threshold also moves. The problem occurs when this threshold moves away from the ideal point in the middle of the two logic states. For instance, if the threshold moves much closer to a logic ‘1’, it needs much more voltage to switch to logic ‘1’ if it was already in a logic ‘0’ state. This corresponds to a degradation in the part’s sensitivity. Conversely, if the threshold is much closer to a logic ‘1’, any noise signal that may not have been large enough to change the state may do so now because of the decreased Signal-to-Noise Ratio.

Examining the individual waveforms, we can see that the Manchester waveform with its regular low-to-high and high-to-low transitions has almost zero baseline wander, making it the ideal data pattern for AC-coupled systems.

The RZ waveform has its baseline wander determined by the number of consecutive zeroes since each of the ones contain two successive transitions. The problem with NRZ is that, depending on encoding protocol, the baseline usually sits between zero and the midpoint giving a worse Signal-to-Noise ratio for the logic ‘0’ condition.

The NRZ waveform has twice the problem that RZ has since the data sits at either extreme for both ones and zeroes. Again, depending on the encoding scheme, the baseline wander moves between both extremes and is the worst of the three cases examined. An additional problem is that the long ones and zeroes strings look more and more like lower-frequency signals presenting the same types of problems listed in the previous section.

An example of the effects on the sensitivity of each of the parts will be covered in the performance evaluation of the NE5211/ NE5217 combination.

Preamplifiers

Each of the preamplifiers that Philips Semiconductors offers is of the transimpedance type. The only difference in each is the available gain and bandwidth of the device. This lets the user mix and match preamplifiers and postamplifiers depending on his application. For a broader summary of the preamplifiers, see references 1 and 2.

NE5212

The NE5212 is the workhorse of the HiFI-100 preamplifiers. With 14kΩ of differential transresistance and 140MHz bandwidth, the NE5212 is suitable for most general purpose applications. With a low input noise current density of 2.5pA/$\sqrt{Hz}$, it offers reliable operation with most photodiodes. The NE5212 is offered in 8-pin plastic, ceramic, and surface-mount packages. A more detailed description of the NE5212 is offered in reference 3 listed at the end of the article.

NE5211

The NE5211, with 28kΩ transresistance offers the highest sensitivity of any of the preamplifiers. It also has the lowest noise at 1.8pA/$\sqrt{Hz}$. Because of the high gain, however, this part has the smallest dynamic range of the preamplifiers with a maximum input current of ±60µA.

NE5210

The NE5210 has the lowest transresistance, 7kΩ differential, of any of the preamplifiers, but it has the highest bandwidth of any preamplifier at 280MHz. Because of the lower transresistance, it has the largest dynamic range of the preamplifiers at ±240µA maximum input.

Postamplifiers

The function of the postamplifier is to take the small signals put out by the preamplifier and to square them up to proper levels so they can be interfaced with TTL. The postamplifier output is then usually sent to a clock recovery and data retiming section . Both postamplifiers in the HiFI-100 chip set, the NE5214 and NE5217, are designed for a minimum bandwidth of 60MHz and typical gain of about 60dB. The data sheet, however, is specified in two ways: one for the minimum sensitivity corresponding to a specific BER, and one for the minimum signal required to trigger the threshold according to certain combinations of R_{THRESH} and R_{HYST}.

The postamplifiers also provide a Signal Detect function, FLAG, that is also TTL compatible. When FLAG is HIGH it means that a signal below the preset threshold has been detected. When FLAG is LOW, it means that the signal detected is above the preset threshold level and is likely to be valid data.

Another function designed to work in conjunction with the FLAG is the JAM. The function of the JAM is to disable the forward path of the amplifier. This is important when the output is supposed to be quiet for signals below the threshold level. This is accomplished by connecting the FLAG output to the JAM input (TTL-compatible). If left unconnected, the JAM will float HIGH, disabling the amplifier's forward path. Therefore, if the FLAG/JAM combination is not desired, the JAM should be tied to ground.

The postamplifier also has an LED pin which offers a visual verification for the FLAG function. By tying the cathode of an LED to this pin through a current limiting resistor (100Ω is fine), you will get the following correspondence: FLAG(HIGH) = LED OFF and FLAG(LOW) = LED ON. Therefore, the LED OFF state represents a noisy condition at the postamp input.

For additional details on the functions in each postamplifier or to find out more detailed specifications, see Reference 5 and each respective data sheet.

NE5214 (Figure 5)

The NE5214 is a postamplifier designed for RZ or Manchester encoded operation. It has a typical bandwidth of 75MHz and comes in an SO-20 package. Distinguishing features are an optional interstage filter between input amplifier A1 and gated amplifier A2. If the filter is not used, Pins 13 and 14 may be shorted together along with Pins 15 and 16.

NE5217 (Figure 6)

The NE5217 is similar to the NE5214 except that it is designed for NRZ operation. The major difference between the NE5214 and NE5217 is that there is no interstage filter between A1 and A2, but the outputs of A2 and the inputs of the Schmitt Trigger A8 are pinned out so that coupling capacitors may be connected between A2 and A8. The coupling capacitors act like a differentiator, passing only the transitions of the incoming data. The Schmitt Trigger has 400mV of hysteresis to insure that the output comparator doesn't change state on noise spikes. The functions of the FLAG and JAM are unchanged.

Considerations for External Components

In our applications, we recommend using a100pF capacitor across the inputs as a low-pass filter to prevent high frequency noise signals from passing through to the output. This capacitor makes a significant (3dB min) improvement in sensitivity.

Auto-Zero Capacitor

The Auto-Zero Capacitor sets the low frequency pole for the input signal. This pole is determined by C_{AZ} as follows: $f_{-3dB} = 640 \times (2 \times \pi \times 1.6k\Omega \times C_{AZ})^{-1}$. The lowest frequency transmitted should be at least 10 times larger than this frequency if it is to be passed through the amplifier. Otherwise, it may be zeroed out by the Auto-Zero loop which functions as a DC to low-frequency feedback loop. For proper operation, the C_{AZ} must be large compared to the coupling capacitors C_{14} and C_{15}.

Coupling Capacitors C_{14} and C_{15} (NE5217)

The optimum value for capacitors C_{14} and C_{15} is 18pF. This has been verified by experiment. For minimal functionality of the part, the ratio between the Auto-Zero capacitor and the coupling capacitors should be a minimum of 250:1. With a 0.1μF capacitor and 18pF, this ratio is about 5500. For better performance, this ratio should be increased by increasing the Auto-Zero capacitor, not shrinking the coupling capacitor.

R_{THRESH} and R_{HYST}

These resistors set the threshold for the FLAG function and the amount of hysteresis built into FLAG function (not to be confused with the hysteresis of the Schmitt Trigger in the forward path). To find the appropriate resistor for your application, refer to the charts given in Figure 7.

R_{PEAK} and C_{PEAK}

R_{PEAK} and C_{PEAK} set the time constant used to determine how long it takes before the FLAG changes from a HIGH (signal absent) to a LOW (signal present). It comes from the following facts:

1.) C_{PEAK} (connected between pin and ground) is in parallel with an internal capacitor of 10pF.

2.) The time constant is proportional to the slew rate specified by $dV/dt = I_{PEAK}/C_{TOT}$, where $C_{TOT} = C_{PEAK} \| 10pF$.

3) I_{PEAK} is set by R_{PEAK} (connected between pin and V_{CC}) by the following formula: $I_{PEAK} = (V_{CC} - 0.8V) / (67.7k\Omega \| R_{PEAK})$.

Performance of the NE5211/5217 Receiver Combination

To verify some of the ideas presented in the system considerations section, one receiver combination, the HiFI-100C which consists of the NE5211 preamplifier along with the NE5217 postamplifier, had its sensitivity tested over several conditions. Variations were made with respect to frequency, data pattern, and external components. In each case, the sensitivity was measured such that the BER rate at that level of input power was 10^{-9}. The receiver tested used a Philips BPF-31 850nm photodetector as an optical front-end. A schematic of this combination is shown in Figure 8. The photodiode was mounted in an SMA-female connector, 2.5mm ferrule connector. The optical cables used to connect to the receiver were 62.5/125μm core/cladding multi-mode fiber terminated in an SMA-male connector.

CAUTION!

This board was designed for good isolation between V_{CC} and Ground to avoid feedback loops and potential oscillations for zero input signal conditions (optical cable not connected). To facilitate this, the top or component side of the board is surrounded by a ground plane. On the bottom side of the board are three V_{CC} sections: one for the photodiode and preamplifier, one for the analog section of the postamplifier, and one for the digital section of the postamplifier.

The danger posed by this situation is that if you wanted to probe the board for various waveforms, you would probably use a probe with an attached alligator-style ground clip. Your first impulse would be to attach this clip to the top and bottom of the board. DON'T DO THIS! This will short ground to any of the three V_{CC} sections and is likely to destroy either the photodiode or the NE5211 . If you need to do any probing, attach the ground clip to the ground posts on the top of the board. Of course, this is not a concern if you have one of the laminated boards.

Measurement Considerations

To make sure that we were measuring the forward amplification path of the postamplifier and were not shutting it down for specific threshold voltages, the JAM was disconnected from the FLAG and grounded to insure that the forward path was always active.

Data Patterns

We tested the receiver with three patterns: a square wave, a PRBS of 2^7-1, and a Pseudo Random Bit Sequence (PRBS) of 2^{23}-1 . Each signal transmitted was in NRZ mode so that for a 50MHz clock signal, the maximum data rate transmitted would be 25MHz. A square wave was used to test the zero baseline wander case and to approximate Manchester code. PRBS=2^7-1 was used as a moderate test of baseline wander, since there could be no more than 7 consecutive ones or zeroes. PRBS=2^{23}-1 was used to approximate some extreme cases of baseline wander by having a maximum of 23 consecutive ones or 22 consecutive zeroes. Harsher baseline wander conditions would occur by using burst-type signals with greater amounts of DC in the signal. Burst data will be discussed in more detail later.

Frequency

The frequencies tested were f_{CLOCK} = 1, 2, 4, 8, 10, 20, 40, 80,100, and 200MHz. Since these were transmitted via NRZ, they correspond to a maximum f_{DATA} = 0.5,1, 2, 4, 8,10, 20, 40, 80, and 100Mb/s respectively. For the minimum frequency signal in each case, take the corresponding f_{DATA} and divide by 7 or 23 depending on which PRBS is used. Of course, for a square wave there is only one frequency transmitted, f_{DATA}.

Sensitivity

For sensitivity values to be meaningful, optical power must be converted to input current for the preamplifier and then into differential volts for the postamplifier. To do this, certain parameters are needed. First of all, the photodiodes responsivity must be considered. As an example, find the voltage input to the postamplifier that corresponds to an input optical power of -33dBm. To find the power in watts, we realize, by definition, that -33dBm is equal to 10 x log (P_{OUT}/P_{IN}) where P_{IN} is 1mW. This gives us P_{OUT} = 500nW. The photodiode's responsivity is R = 0.35Amps/Watt. So multiplying through we get the input current to the preamplifier I_{IN} = P_{OUT} x R = 175nA. To find what differential voltage at the output of the preamplifier is, we can multiply the input current by the differential transimpedance of the NE5211, TZ = 28kΩ. So, V_{DIFF} = I_{IN} x TZ = 4.9mV_{P-P}. This is just a typical value calculation. To find the full range of sensitivity, temperature and voltage variations must be considered.

Bit-Error Rate

Another thing to be considered is how does this input signal compare with the Bit-Error Rate? For the NE5211, the input noise current is 1.8pA/$Hz^{1/2}$. To find out what the Signal-to-Noise Ratio (SNR) is,we have to find what the actual noise current is based upon the density. Suppose that the signal measured in the previous paragraph was taken at 60MHz. The square root of this gives 7.75x$10^3$$Hz^{1/2}$. Multiplying the two together gives a noise current of 13.9nA. The SNR would then be 138nA/ 13.9nA = 12.6 or 22.0dB. Using a chart in Figure 9 from reference 4, we note that this roughly corresponds to a BER between 10^{-6} and 10^{-7}. So, we can conclude that for a desired BER of 10^{-9}, the power level is too low. You can also work backwards, finding the minimum power level needed for a specified BER.

For accuracy, each measurement was tested with at least 3x10^{10} bits. For a desired BER of 10^{-9}, this gives us an accuracy of (1 - [(10^{-9}) x (3x10^{10})]$^{-1}$) x 100 = 96.7%. Of course, for more accuracy more bits could be sent but the time needed to test the receiver would grow accordingly.

Results

Figure 10 shows the results for board A, where the Auto-Zero capacitor to coupling capacitor ratio is 1:1 . This is clearly the worst case. The square wave gives us the best sensitivity because it has the least amount of baseline wander. The next best sensitivity is for PRBS=2^7-1 because of the seven consecutive ones and zeroes in the pattern. The worst case is for PRBS = 2^{23}-1 which has lower frequency components which are not even passed for values lower than 8MHz regardless of the strength of the input. As expected for the pseudo-random sequences, the sensitivity goes down as the frequency goes down, as the low frequency components of the PRBS data stream are cancelled out by the Auto-Zero loop.

Figure 11 shows case B where the capacitor ratio has increased to 1000. We get significant improvements for the PRBS = 2^{23}-1 case, but not too much more for the other two data patterns.

Figure 12 shows case C which has increased the ratio even more, to 10,000, by increasing C_{AZ} to 1µF. This improves the performance of PRBS = 2^{23}-1 even more. It should be noted that the sensitivity in all cases (A, B, and C) goes down beyond 80MHz because of the natural -3dB roll-off of the postamplifier. It appears that the best operation is around 20MHz for these particular external components.

Figure 13 shows how boards A, B, and C performed for a square-wave input. Judging by the y-axis scale, there is no appreciable difference in either of the cases for this type of input.

Figure 14 shows how boards A, B, and C performed for an input of PRBS = 2^7-1 . C offers the best performance at lower frequencies.

Figure 15 shows an interesting comparison of how board C (with C_{14} = C_{15} =100pF) performs compared with the capacitors shorted (the equivalent of the NE5214). Board C is consistently 2 to 3dB better at frequencies lower than 20MHz.

Figure 16 shows the performance of boards A, B, and C along with board C using 18pF coupling capacitors for an input of PRBS = 2^{23}-1. The 18pF value performs better at more points than the 100pF value and is thus chosen as our optimum coupling capacitor value. (Values lower than 18pF actually degraded the sensitivity.)

Burst Data Transmission

One of the most difficult types of data to transmit is so-called "burst" or "bursty" data. Burst data can best be described as a pulse or series of pulses followed by a long period of no transitions. The problem caused by burst data is clearly the lack of transitions in the data stream. For DC-coupled, Auto-Zero loops, a forced high or low voltage would not cause problems. In DC-coupled, Auto-Zero systems, as the length of the not transition or "dead time" increases, the likelihood that the signal will be canceled out by the Auto-Zero loop increases. This will result in bit side errors being transmitted. In AC-coupled systems, this causes a drift to the DC-bias levels on either side of the capacitor, usually resulting once again in bit errors. It is for this reason that many encoding schemes require a minimum number of transitions per parallel word.

However, other types of faults may look like burst data transmission which should be recognized as some type of error condition. For instance, if the link is cut off for some reason and no data gets through, should this be recognized as a zero or a one? In an AC-coupled system, how would the receiver distinguish between a broken link which balances the inputs versus a long series of ones or zeroes? More often than not, a coding scheme designed to flag after a certain amount of time of no transitions or a SIGNAL DETECT function that would respond only to DC signals would have to be implemented.

One receiver, an NE5210/NE5217 combination was tested under burst-data conditions. Two data patterns were used. The first pattern used had twelve 40ns pulses or 11 and a half cycles of 25MHz square wave cycled over 2048 pulses of the same size (2048x40ns = 81.92µs for a 12.207kHz repetition rate). The duty

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cycle for the pulse burst is (12x2/pulses)/2048 pulses = 1.17% or 11,719ppm. This input signal burst is shown in Figure 17 (top trace). The output trace (output of NE5217) is shown in the bottom trace. In this configuration, the FLAG output is connected to the JAM input. This presents a problem.

Because the JAM function is implemented at the A2 amplifier and not at the gate, a charge is stored on the coupling capacitors to the Schmitt Trigger, C14 and C15 "freezing" the output in the logic '1' state for 100ns before it changes state to zero, 60ns longer than it should be.

To see if this problem is at all related to the number of pulses transmitted or to the burst repetition rate, a worst-case testing was attempted by sending a single 40ns pulse at a repetition rate of 24 pulses = 24x40ns = 960ns or 1.04MHz. This gave us a duty cycle of 1/24 = 4.17%. The output is shown in Figure 18. The JAM is still connected to FLAG and the FLAG signal is shown in the top trace. Note that the part can only transmit through the A2 amplifier when FLAG = JAM is low. During the no-transition periods, the part considers it to be a loss of signal. This would be the same case as if the input were cut off as a result of a break in the link. The middle trace has the TX input and the bottom trace shows the pulse stretched output similar to the previous output trace.

To check performance independent of the state of FLAG/JAM, we tried the same thing but with the JAM grounded. In Figure 19, we see that JAM grounded in the top trace. In the middle trace we have the input signal. It is inverted in this instance because of a trigger change. The output of the NE5217 is shown in the bottom trace with no evidence of the pulse stretching. To verify the performance, we took this signal and measured it to a sensitivity of -24.5dBm optical with 850nm light to a BER of $\leq 10^{-9}$ (measured with a minimum of 3×10^{10} bits). If we used an NE5211 we might expect performance around -30dBm to -32dBm.

If transmission of burst data is a necessity and the FLAG function is essential, you could JAM the part externally with an OR gate. When no signal/no transition is the case, the FLAG would be high and the output would always be HIGH. With the FLAG LOW (signal present), you would get whatever the NE5217 output is. This is shown in Figure 20.

REFERENCES

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2 . "Optical Fiber Communications", by John M. Senior, Prentice-Hall International Series in Optoelectronics, 1985.

3. "A Low Cost 100MBaud Fiber Optic Receiver", by William D. Mack, Robert G. Meyer, and Ki. Y. Suh, Philips Semiconductors Company, presented at FOC/LAN (Fiber Optic Communications/ Local Area Networks), October 1987.

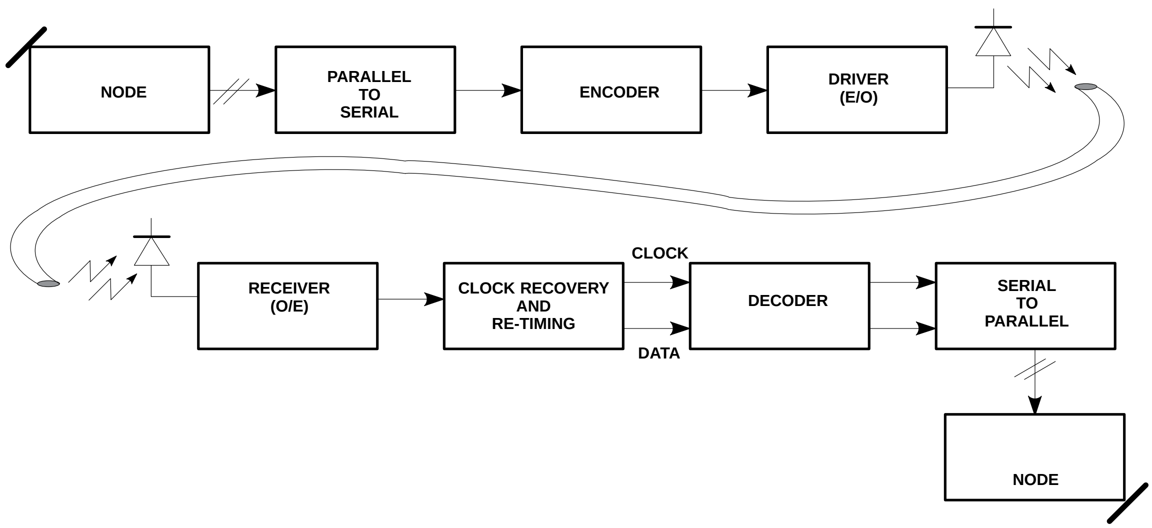

Figure 1.

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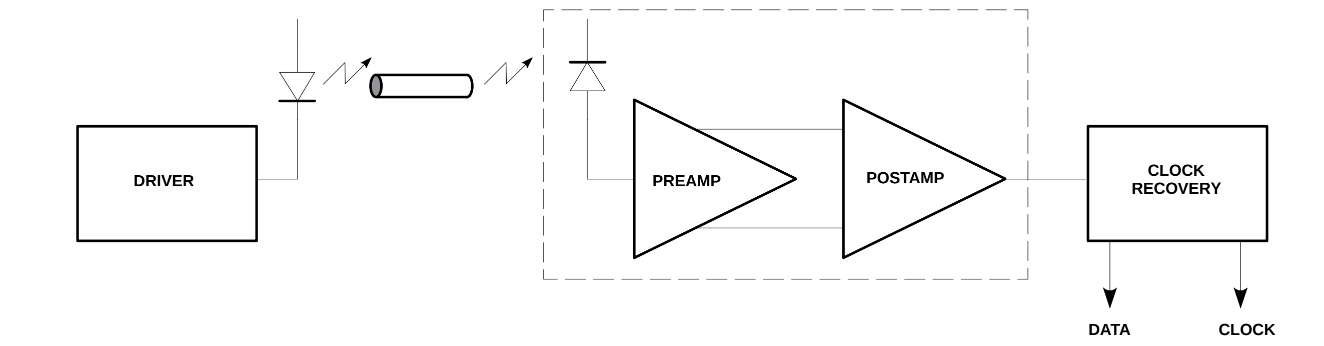

Figure 2.

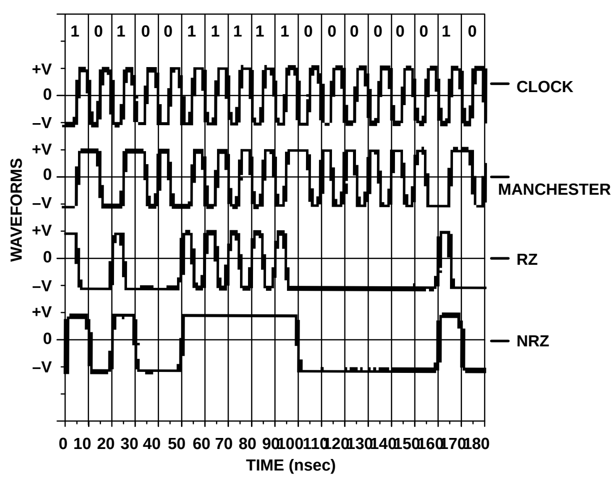

Figure 3. Various Codes for Same Data

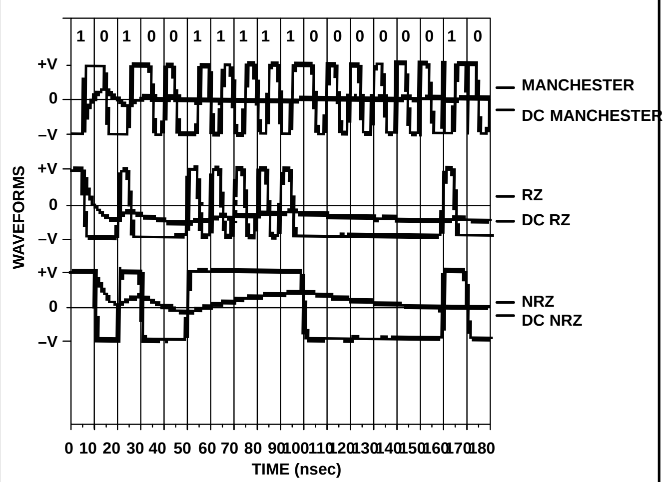

Figure 4. Different Coding Schemes with Associated DC Averages; Baseline Wander is from 0.

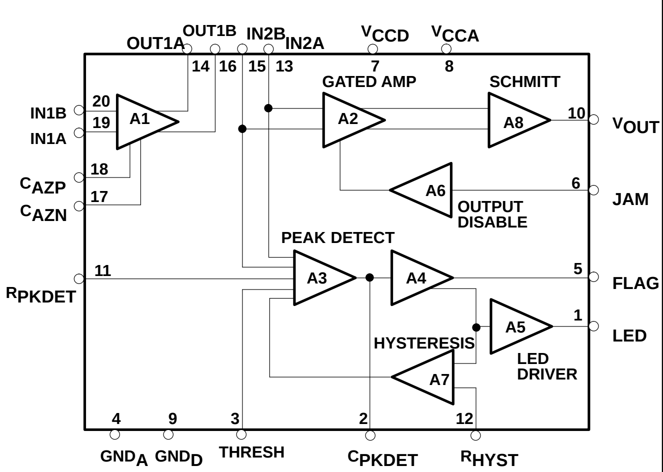

Figure 5. NE5214 Postamplifier Block Diagram

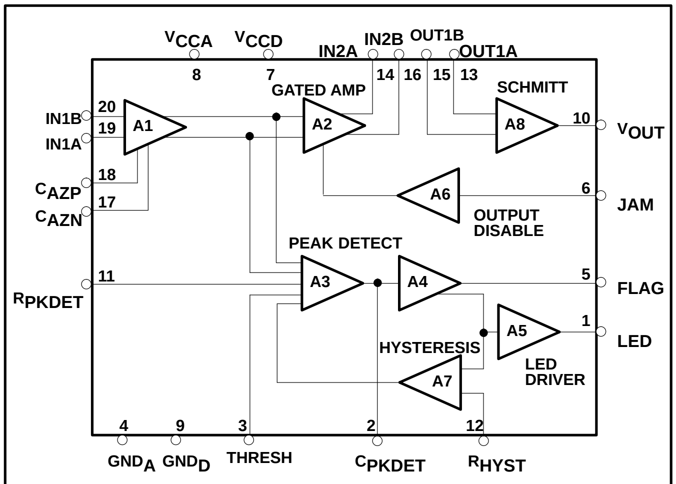

Figure 6. NE5217 Postamplifier Block Diagram

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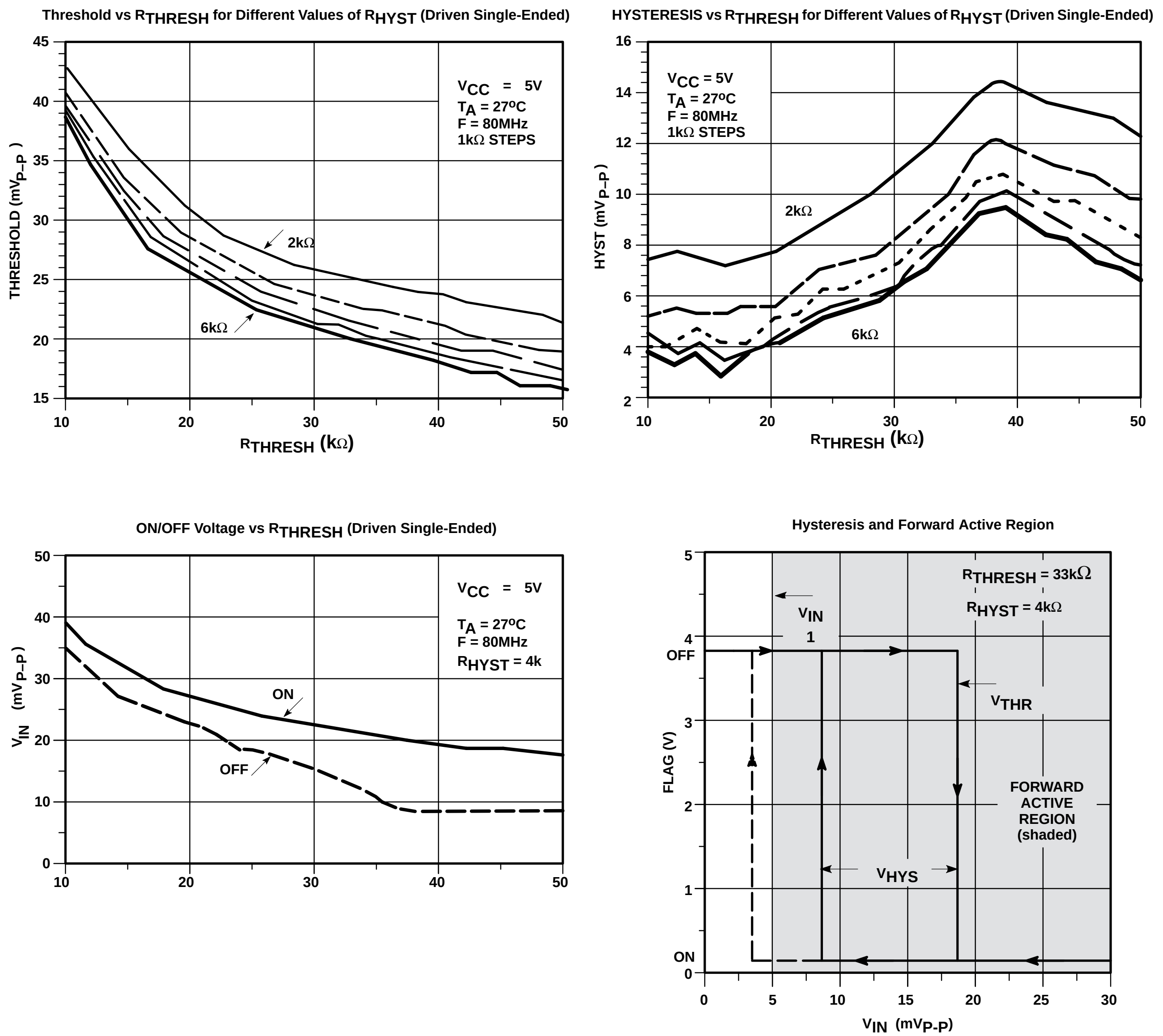
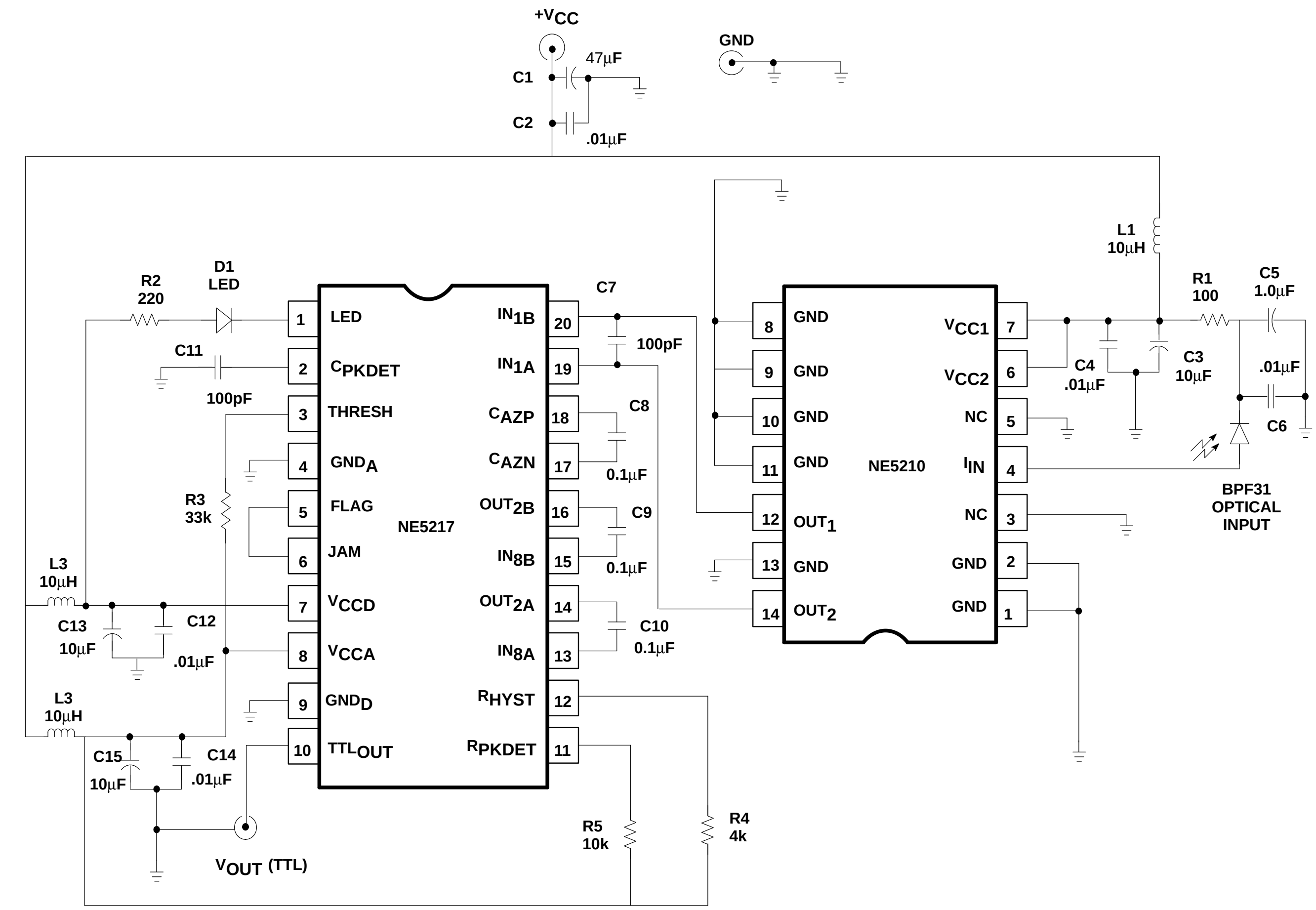


Figure 7. Typical Performance Characteristics



NOTE:
THE NE5210/NE5217 combination can operate at data rates in excess of 100Mb/s NRZ

Figure 8. A 50Mb/s Fiber Optic Receiver

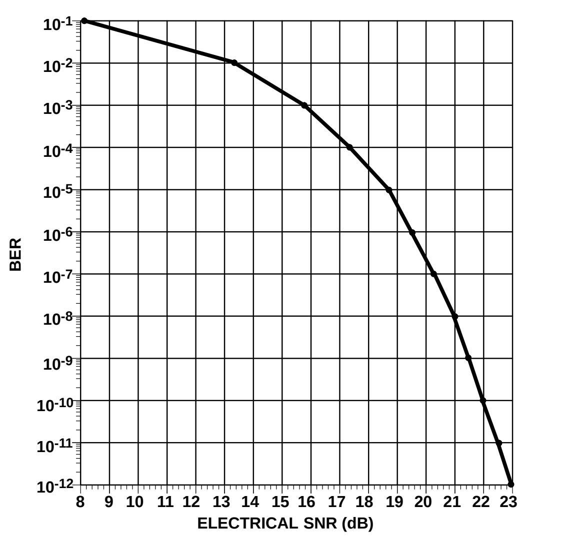


Figure 9. Bit-Error Rate vs Electrical Signal-to-Noise Ratio

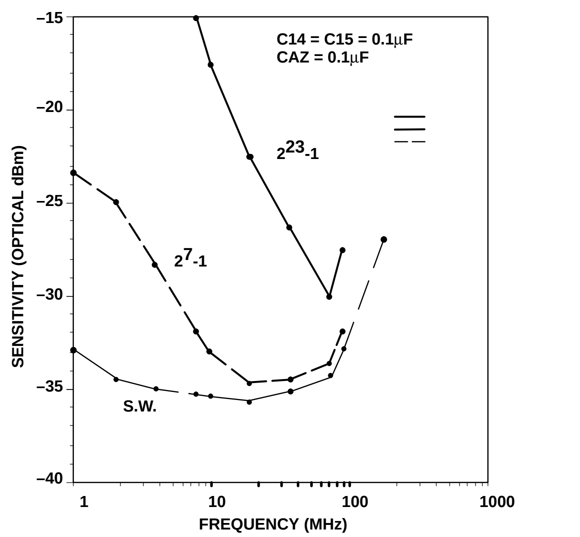


Figure 10. NE5211/17 Sensitivity vs Frequency for Square Wave, 2^7-1, 2^{23}-1 Patterns

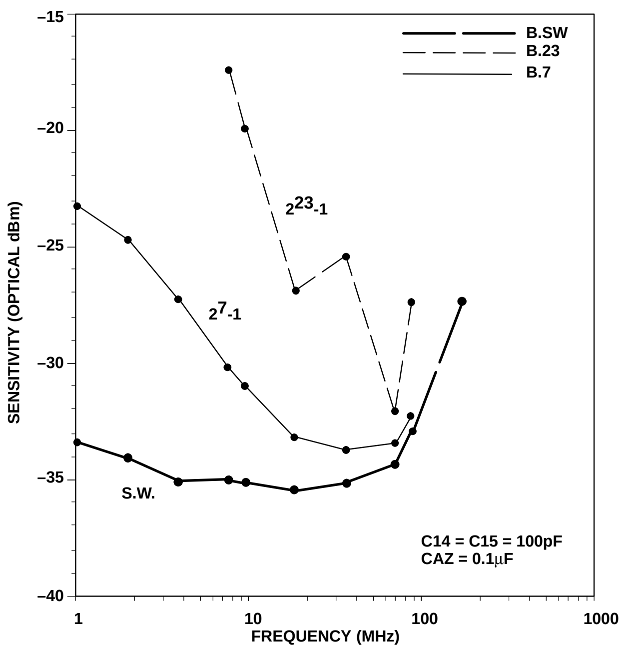


Figure 11. NE5211/17 Sensitivity vs Frequency for Square Wave, 2^7-1, 2^{23}-1 Patterns

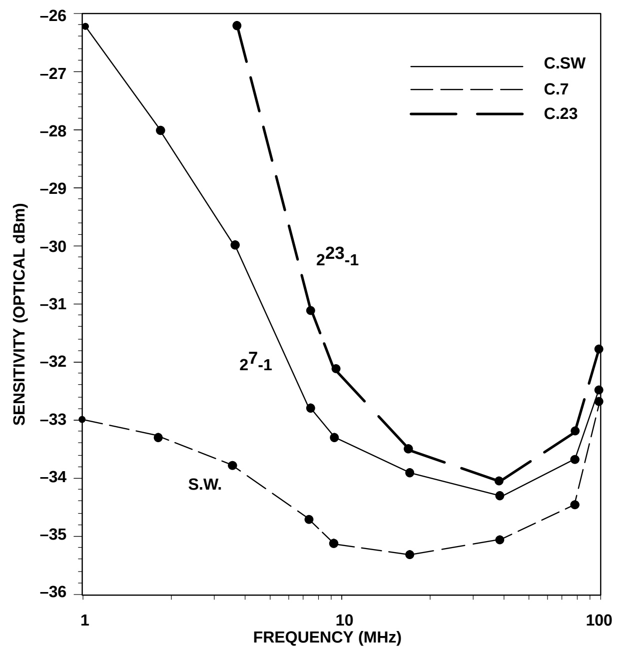


Figure 12. NE5211/17 Sensitivity vs Frequency for Square Wave, 2^7-1, 2^{23}-1 Patterns

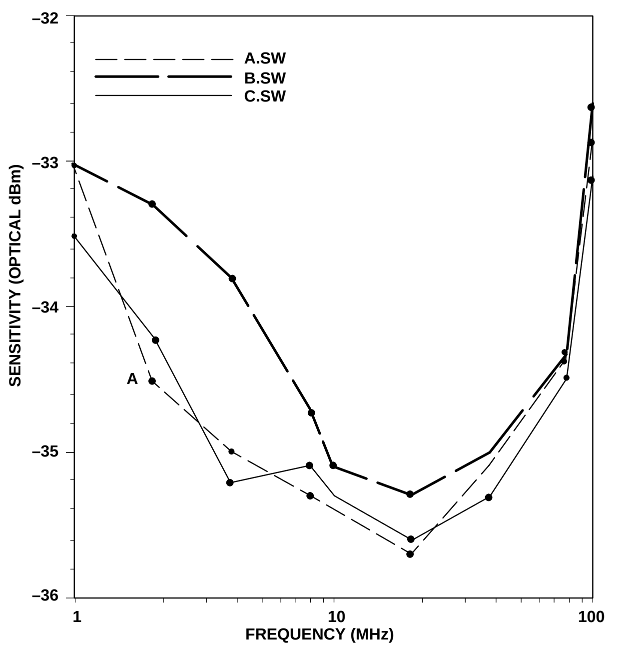


Figure 13. Optical Sensitivity versus Frequency for Square Wave Input (Configurations A, B, C)

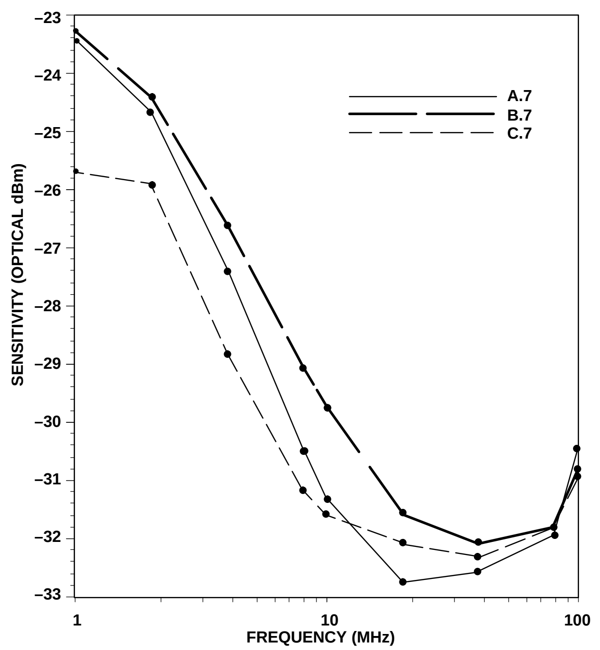


Figure 14. NE5211/17 Sensitivity vs Frequency for PRBS = 2^7-1, Input (Configurations A, B, C)

Low cost, TTL fiber optic receivers for up to 100Mb/s NRZ AN1443

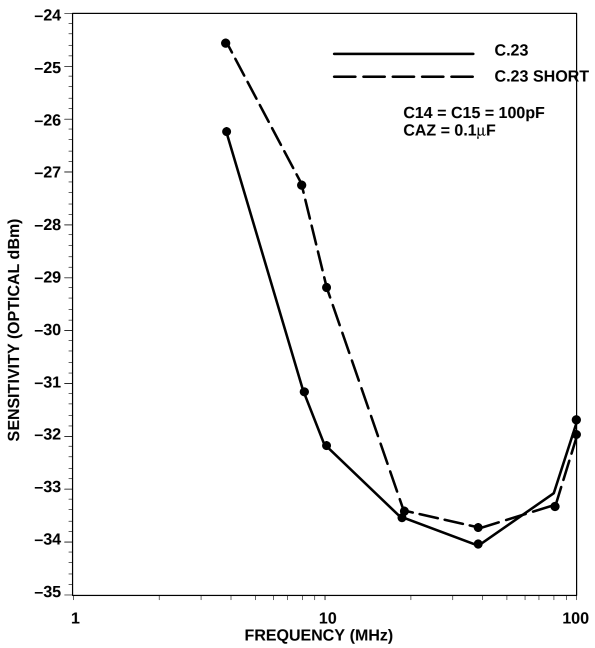


Figure 15. NE5211/17 Sensitivity vs Frequency for PRBS = 2^{23}-1

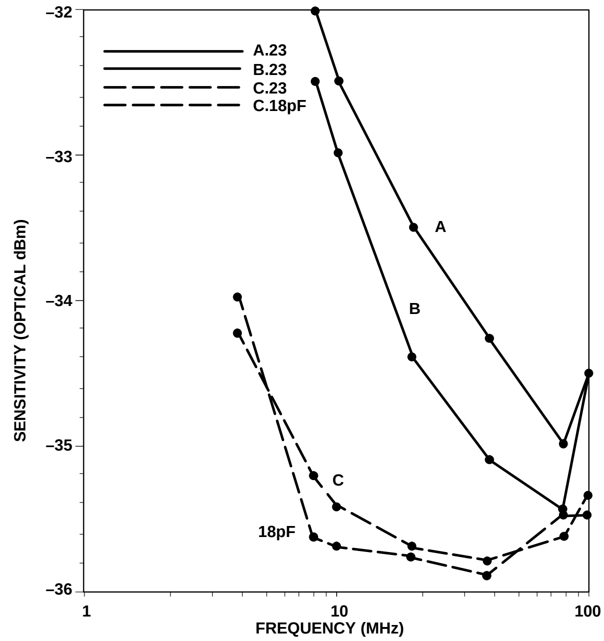


Figure 16. Optical Sensitivity versus FrequencY PRBS = 2^{23}-1 (Configurations A, B, C-100pF, C-18pF)

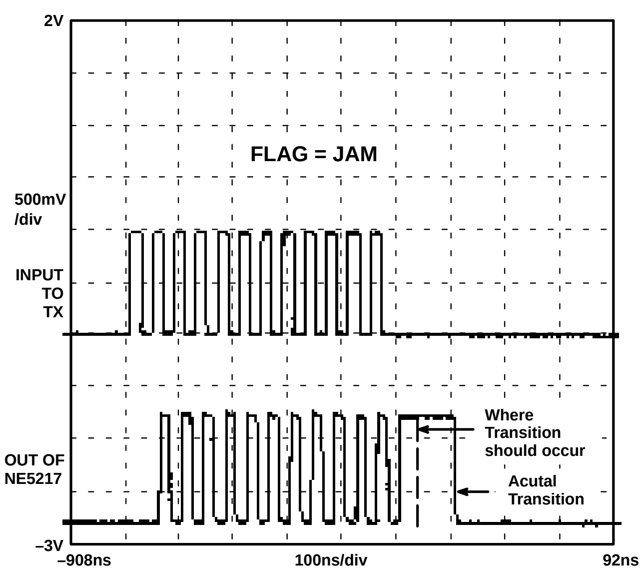


Figure 17.

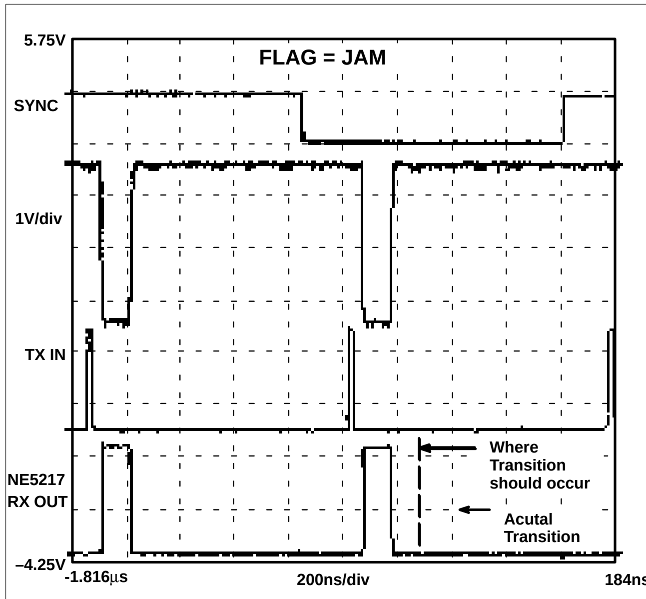


Figure 18.

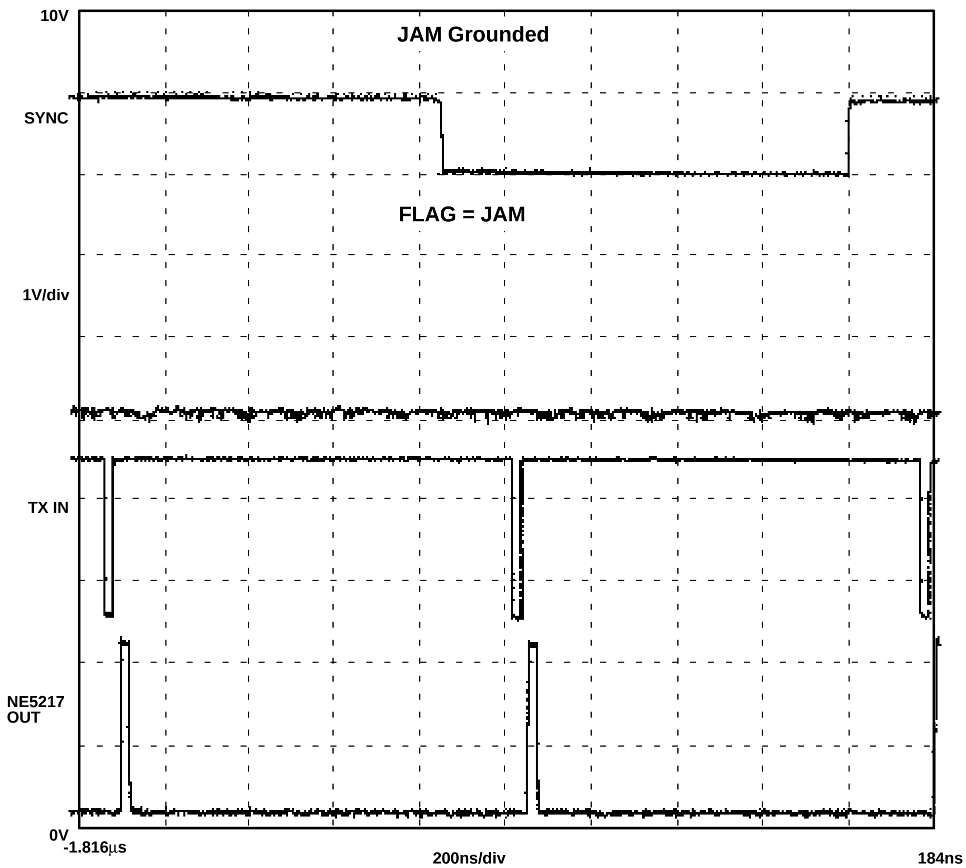
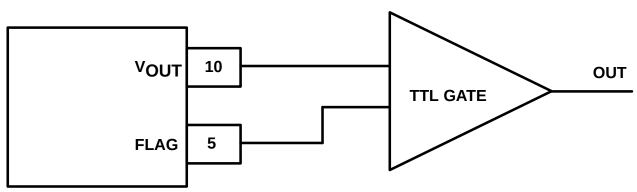


Figure 19.



	V_{OUT}	FLAG	OUT
SIGNAL PRESENT	0	0	0
	1	0	1
SIGNAL ABSENT	0	1	1
	1	1	1

Figure 20. Externally Jamming Part