Using the F100250 for Copper Wire Data Communications

National Semiconductor
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Ideal "digital" signals do not exist, especially when the signal must travel from source to destination over any current-carrying conductor. The world of "digital" signals is truly the world of high-frequency analog and radio-frequency (RF) amplifiers and energy transmission systems. This is especially true for the case of high-speed copper wire data communications networks. The problems of sending signals over the wire interface, whether a printed circuit board or a coaxial cable, require a knowledge of transmission line theory.

To effectively use devices like the F100250 Line Transceiver in high-speed data communications networks, the system designer needs to be acquainted with several subjects among which are: the effects of using pulse excitation on a transmission line, a knowledge of the various forms and modes of data-transmission-line circuit operation, familiarity with the problems of working with long transmission lines, a working knowledge of the driver and receiver, their electrical characteristics, and where and how to use them. This applications note cannot treat the whole subject of "digital" data transmission since the scope of that subject could and has filled whole volumes. This note will touch upon the transmission line topic in conjunction with offering helpful suggestions on how to more effectively use the F100250 Line Transceiver.

F100250 DESCRIPTION AND OPERATING FEATURES

The F100250 is a quintuple, differential-line transceiver with the unique capability of being able to transmit and receive differential-mode signals simultaneously on the same transmission line. The F100250 is part of the National Semiconductor F100K ECL family. As such it shares ECL interface signal characteristics in common with the F100K family.

The circuit of the F100250 *(Figure 1)* is comprised of a line transmitter with differential output, a differential receiver with transparent latch, signal separation circuitry, and internal line termination circuitry. The transmitter is a single-ended input to differential-output amplifier which connects to the line through an active-resistive bridge network. This network provides the correct driving-point and termination impedances for the transmission line and forms part of the received-signal separation circuitry.

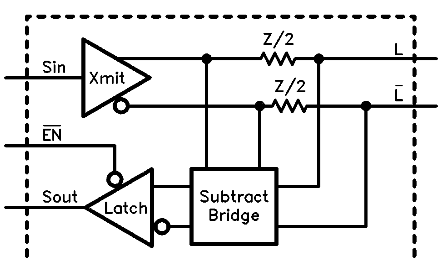

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FIGURE 1. F100250 Block Diagram

The receiver consists of a voltage subtractor and hysteresis circuit followed by an amplifier and emitter-follower output. The receiver has a common-mode voltage immunity of $\pm 1V$. The possibility of oscillation in response to slow rise or fall times is reduced by using hysteresis; and the ability to detect noisy signals is improved. The typical hysteresis level of the F100250 is 50 mV.

The receiver incorporates a transparent latch for data retention in synchronous-type operations. The level-sensitive, latch $\overline{\text{ENABLE}}$ pin simultaneously controls the operation of all latches in the part. Data present on the line inputs (L and $\overline{L}$) prior to taking $\overline{\text{ENABLE}}$ high is retained in the latch, assuming proper setup and hold timing is met. The latch is fully transparent when $\overline{\text{ENABLE}}$ is low.

Termination is provided internally for 30 AWG twisted-pair lines which have a nominal impedance of 150Ω. Higher or lower impedance values may be accommodated by use of a suitable external termination network.

Bi-Directionality

The hallmark of the F100250 is its ability to simultaneously send and receive differential-mode NRZ signals over the same line. This operational mode is known as "baseband full-duplex". By contrast, full-duplex operation is normally accomplished using frequency-division multiplex (FDM) techniques. A common example of which is the 103 or 212A telephone line modem. The F100250 uses a balanced-bridge to separate the transmitted and received baseband signals.

The F100250 can also operate in a uni-directional manner. It is suggested that the input to the transmitter for the unused direction be held at a fixed mark or space level. The ECL signal inputs (Sin 1–5) incorporate 50 kΩ pull-down resistors for the purpose of holding the unused input at a low (inactive) logic level.

The F100250 cannot operate in a party-line or multi-drop mode due primarily to the fact that the transmitter outputs cannot be turned off (i.e., made high-Z or TRI-STATE®). Also, connecting more than two devices to the line would cause a multiple mismatch to occur since the device's output is self-terminating. This restriction should present few problems since the primary use for the F100250 will be in the highest speed point-to-point type applications.

SUITABLE TYPES OF TRANSMISSION LINES

Twisted-Pair Lines

The F100250 is designed for operation over 150Ω, 30 AWG twisted-pair lines. However, it will operate with a variety of other line types and impedances if an appropriate termination method is used as will be shown in the application example.

Twisted-pair transmission lines *(Figure 2)* are available shielded and unshielded, singly and in multi-pair cables, in popular ribbon configurations and in combinations of these. Impedances range from 93Ω for 32 AWG solid, unshielded, ribbon cable to 124Ω, 25 AWG, multiple, individually-

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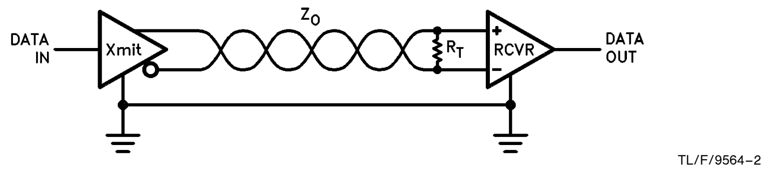


Note: $R_T = Z_0$

FIGURE 2. Terminating Differential Twisted Pair

shielded pair and the aforementioned 150Ω, 30 AWG, unshielded, individual pair. Twisted-pair lines exhibit substantial attenuation and dispersion. This may be seen as the rounding and slowing of fast rise-time signals as they travel down the line. Attenuation can range from 10 dB to over 30 dB per 100 feet at 100 MHz. Propagation velocities range from 0.66 to 0.78 the speed of light (1.3 ns to 1.6 ns/ft). In addition, the electrical characteristics of twisted-pair lines do not permit their analysis by the more traditional methods used for coaxial lines. For this reason, the engineer designing twisted-pair wire transmission systems must be equipped to take and interpret measurements on the particular line type chosen for the application. This is the best, and in some cases the only, way in which the capabilities and limitations of the particular driver-line-receiver combination may be understood. Test equipment and fixturing will be described later in this note to help in making meaningful measurements.

Multiple-Pair Cables

Multiple twisted pair cables are a popular method of interconnecting computing equipment and peripheral devices. Additional considerations must be given when using these types of cables. Some of these are: pair-to-pair coupling and capacitance, pair-to-shield capacitance and pair-to-pair relative propagation delay difference. For example, the propagation delay of a single pair in the cable may be 1.5 ns/lineal-foot and the relative delay between pairs is 0.5 ns/lineal-foot. The design value for the actual per-unit delay would be 1.45 ns to 1.55 ns/lineal-foot. That is, the cable will exhibit a pair-to-pair per-unit-length delta-delay of 0.10 ns. This delta-delay value can be considered additive per unit length between any two given pairs. However, the overall delay value for a length of the cable will be that of the pair with the longest electrical delay. The overall relative delay of the cable will be the difference between those pairs with the longest and shortest electrical lengths. This is an important delay factor to be considered in determining the minimum unit interval (hence, the maximum data rate) which can be propagated by the network.

There is another point to remember when determining the delay of multi-pair cables. The delay figures specified in cable data sheets (when given) are normally expressed as delay-per-unit cable length. This value is greater than the actual delay-per-lineal-foot of the pair itself because the layers of pairs are laid-up in a spiral wrap. The fact that the outer pairs traverse a greater distance due to the spiral wrap usually means that they have greater delay per unit cable length than the inner layers of pairs.

Coaxial Cables

Coaxial cables offer a near ideal transmission medium for "digital" data communications signals. Among their more desirable features are: a high degree of shielding, wide bandwidth capability, high velocities of propagation and low attenuation at high frequencies. Coaxial cables are now available paired and in multiple (or ribbon) configurations. Impedances range from 50Ω to 125Ω for standard coaxial cables and up to 200Ω for twinaxial cables. Appendices A and B give a partial list of available coaxial cable with abbreviated data.

Ribbon coax offers a convenient and compact transmission line with mass-termination connector capability. Normally available in 75Ω and 93Ω versions, it has excellent shielding properties because of its foil shielding. Available types have moderate attenuation and high propagation velocity.

Terminating coaxial cables is easily done with a parallel termination resistor or by terminating each coax individually as shown in *Figure 3*.

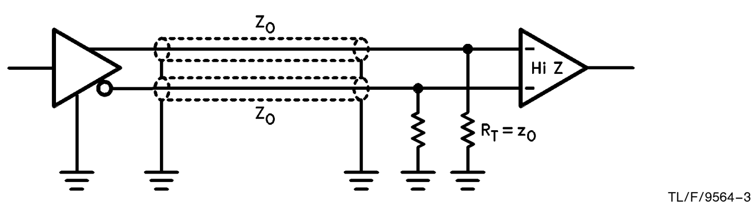


FIGURE 3. Terminating Coaxial Cables

As with twisted-pair cables, the electrical length of both coaxial cables used for the differential pair must be the same. It is usually desirable to temperature-cycle and flex the cable prior to cutting and measuring to electrical length. This will relieve stresses resulting from manufacturing and storage on spools. Without stress-relieving, the cable may change in physical length and therefore in electrical length with unpredictable results for network timing.

Measuring the Electrical Length of Cables

Measurement of the electrical length of cables, coaxial or twisted-pair, can be done using a time-domain reflectometer (TDR) which measures physical length by determining the round-trip delay of a fast rise-time pulse signal. The TDR can also be used to check for defects in the cable such as shorts or opens and impedance discontinuities caused by sharp bends or kinks. However, the accuracy of the TDR, usually 1%, does not allow precise length or timing measurements to be made, especially on very short or long lengths. More precise measurements require the use of multi-frequency phase delay techniques using a vector voltmeter or network analyzer and precision frequency sources.

Transmission Lines on Printed Circuit Boards

Printed circuit wiring may also be used as interconnections for the F100250. Line impedances of 75Ω to 100Ω are easily achieved with conventional manufacturing technologies. The main points to observe when laying out differential networks on printed wiring boards are: that both conductors be the same electrical length and that they are the same impedance. This will insure that no skewing of the differential signal occurs at the receiver input. Skewing appears as an offset in input differential voltage to the receiver.

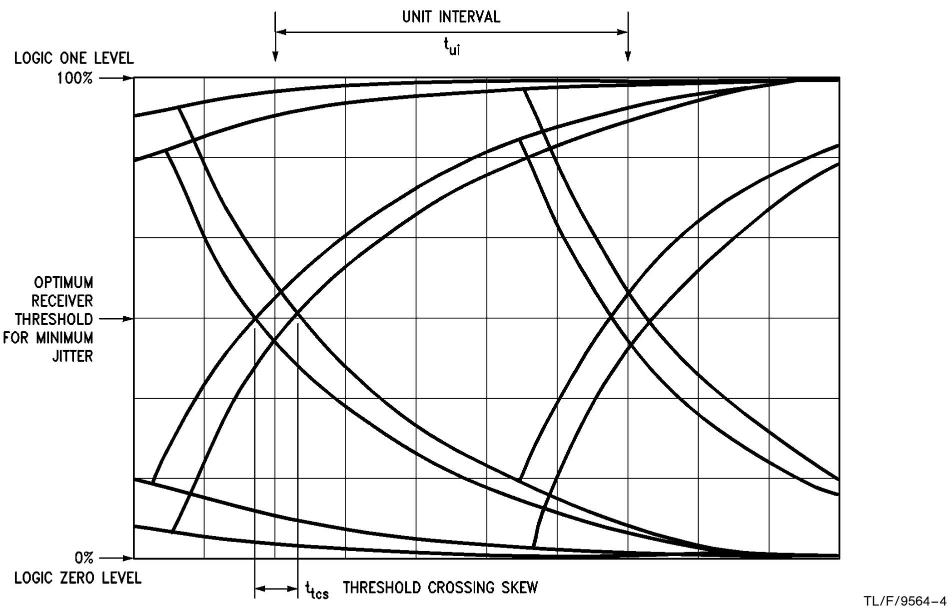
A wealth of information on printed wiring design is contained in the "F100K ECL Databook and Design Guide". Differential techniques are also covered.

Estimating Signal Quality

Before proceeding to the F100250 data transmission system design example, some concepts and terms for the various signal abberations which will be encountered need to be defined.

Signal quality is concerned with the variation between the ideal instants of the original data signal and the actual transition times of the recovered data signal *(Figure 5)*. Recovered data transitions may be displaced in time from their ideal instants. This is caused by a new wave arriving at the receiver before the previous wave has reached its final value. This is termed "intersymbol interference". It can be reduced by making the unit interval of the signal long with respect to the rise (or fall) time of the signal at the receiver input. Reducing the modulation rate for a given line length or vice versa will reduce this form of interference.

Another form of received-signal distortion present with synchronous signalling, like NRZ, is "isochronous distortion". This is the ratio of the unit interval to the absolute value of the maximum measured difference between the actual and theoretical significant instants. In other words, it is the percentage of the unit interval that is peak-to-peak time jitter of the data signal *(Figure 4)*. If the peak-to-peak time jitter of the transition were one-half of the unit interval, the isochronous distortion would be 50%.



$$\text{Peak-to-Peak Jitter} = \frac{t_{tcs}}{t_{ui}} \times 100\%$$

FIGURE 4. Estimating Peak-to-Peak Jitter

"Bias distortion" is the shortening of the duration of the mark-bits with respect to the space-bits or the reverse *(Figure 5)*. It may be caused by a shift of the receiver threshold, asymmetrical driver output levels or both. Together, bias distortion and intersymbol interference are called "systemic distortion". This is because their magnitudes are determined by data transmission system characteristics. Other forms of randomly-occurring distortion such as noise and crosstalk are called "fortuitous distortion". These are due to factors outside the data transmission system.

Signal Quality Measurement

Measurements of signal quality on any transmission system should always be designed to show the effects of intersymbol interference and bias distortion. This means that the test signal must be capable of showing both effects. The use of a simple NRZ dotting pattern (a signal with 50% duty cycle and frequency of one-half bit-time) cannot show intersymbol interference due to its symmetry. It can show bias distortion as a change in duty cycle for the recovered dotting pattern.

The use of a random NRZ data pattern can show both types of interference by its unpredictable bit sequence. A pseudo-random NRZ data generator built from standard F100K ECL devices is shown in *Figure 6*. This circuit is capable of producing a random sequence (2E20)−1 bits in length at frequencies up to about 240 MHz. When the data produced by this circuit is transmitted over the line and viewed on a suitable oscilloscope, a so-called "binary eye pattern" will be seen. This pattern results from the superposition of alternating mark and space bits during each unit interval. It is so called because the pattern's center resembles an eye.

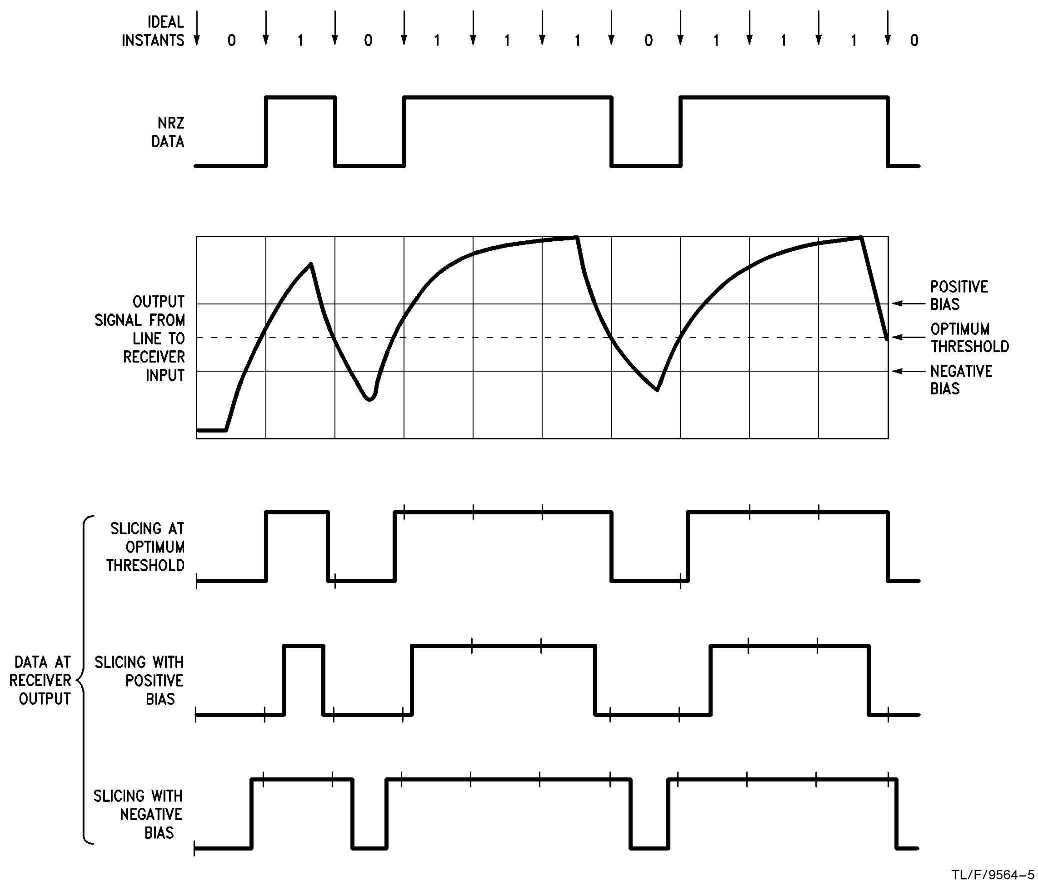
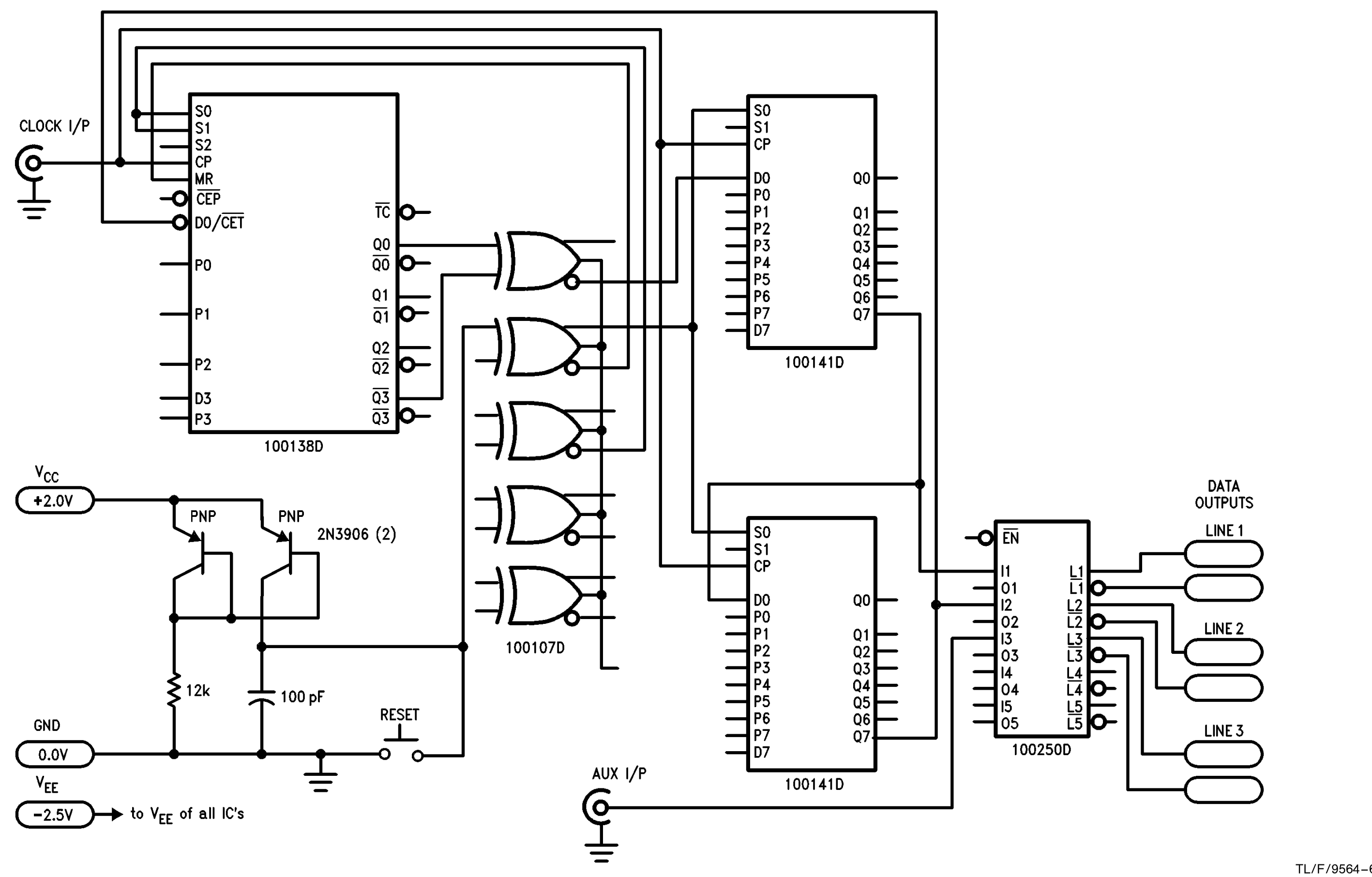


FIGURE 5. Bias Distortion



Note 1: Termination resistors not shown.

Note 2: Split supplies for test equipment convenience.

Note 3: Clock input 230 MHz max.

FIGURE 6. F100250 Differential Line Test Circuit

The eye-pattern is a useful tool to measure data signal quality *(Figure 7)*. The spread of transitions crossing the receiver input threshold can be used as a direct measure of isochronous distortion (peak-to-peak jitter). Rise and fall time can be measured by using the self-references of 0% and 100% resulting from the long sequence of mark and space bits.

The noise margin of the system can be measured as the height of the trace above or below the receiver threshold level at the sampling instant. The eye-pattern can even be used to determine the characteristic impedance of the transmission line. The method is discussed in Appendix C.

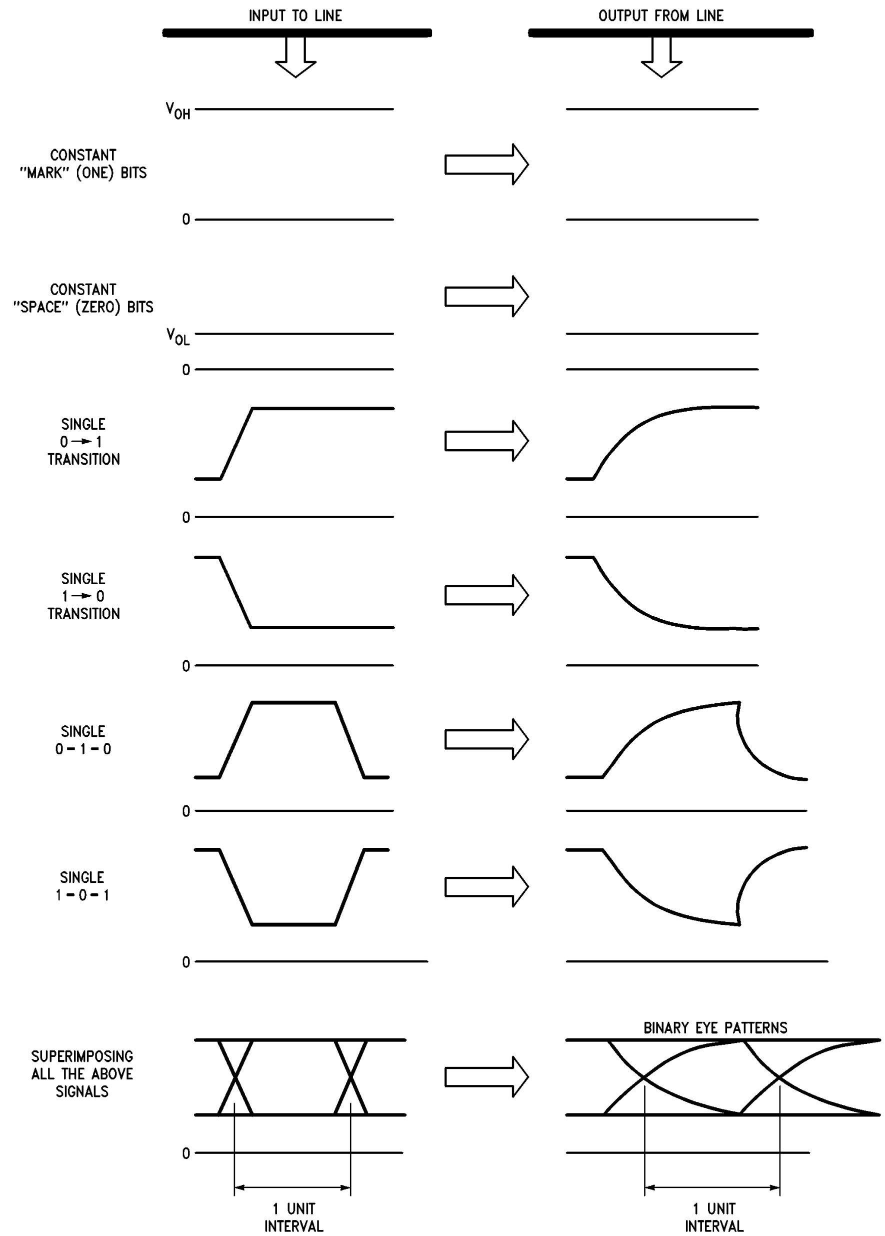


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FIGURE 7. Formation of the Binary Eye Pattern

The eye-pattern gives, in some ways, the minimum peak-to-peak transition jitter for a given line length, type, pulse code, and modulation rate. This is because the pattern results from intersymbol interference and reflections (if present). Minimum jitter conditions only result if: 1) the mark and space signal levels from the driver are symmetrical and the receiver's threshold is set at the mid-point of these levels; 2) the line is terminated in its characteristic impedance; and 3) propagation delays through both transmitter and receiver for both logic states is symmetrical and without relative skew. Signal quality is reduced if any of these conditions is not met.

The decision threshold shown by the displayed eye-pattern for a particular driver and modulation rate shows the effects of receiver bias (or threshold ambiguity) and offset. The slope is small in the threshold region for signals with greater than 20% isochronous distortion. Therefore, small amounts of bias produce large increases in isochronous distortion. A good practice is to design systems to have less than 5% transition spread as shown in the eye-pattern. This minimizes the effects due to bias and simplifies design requirements for line transmitters and receivers.

Application Example—
Putting the F100250 to Work

The F100250 excells at transmitting over twisted pair wiring. *Figure 8* shows an example using 50 meters of 106Ω, 26 AWG, unshielded pair. The pair used is one of 25 in a multi-pair cable specifically designed for digital signal transmission. Note that termination resistors have been added for improved impedance matching to the F100250 line terminals. The scope photos in *Figures 9a* and *9b* show the composite signal conditions at the receiver input. Two pseudo-random signals, 10 MHz and 50 MHz, are present on the line in this example.

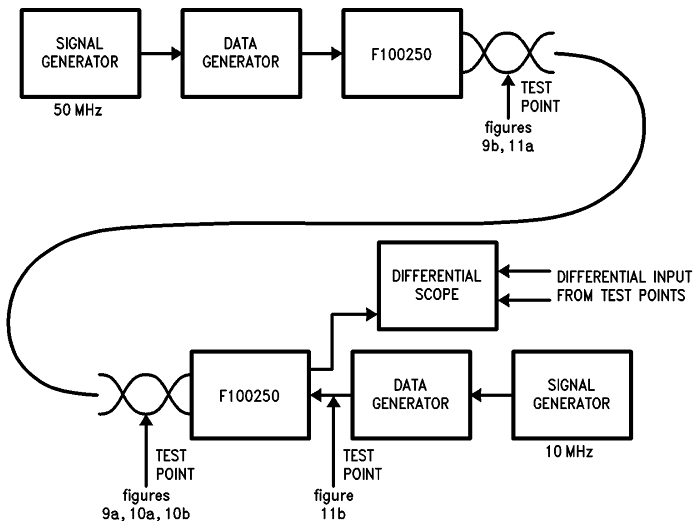

FIGURE 8. Twisted Pair Test Circuit

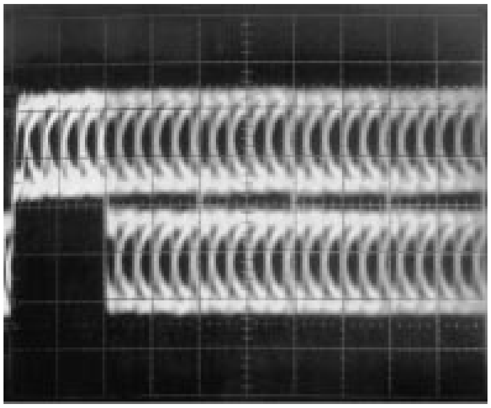

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FIGURE 9a. Composite Signal at Receiver Input (50 MHz) Showing Both 10 MHz and 50 MHz Signals on the Line (Differential Mode)

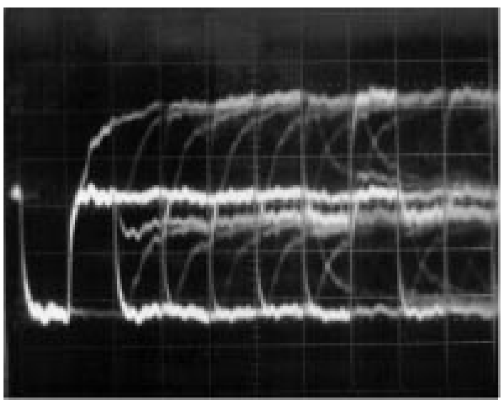

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FIGURE 9b. Composite Signal at Transmitter Output (50 MHz) Showing Both 10 MHz and 50 MHz Signals on the Line (Differential Mode)

Figure 9a shows the composite signal at the receiver input for the 50 MHz signal. Note that both signals appear with the received signal (50 MHz) ''riding on'' the locally transmitted signal (10 MHz). *Figure 9b* shows the 50 MHz signal as transmitted. Signal attenuation is approximately 2 dB for the 50 MHz signal.

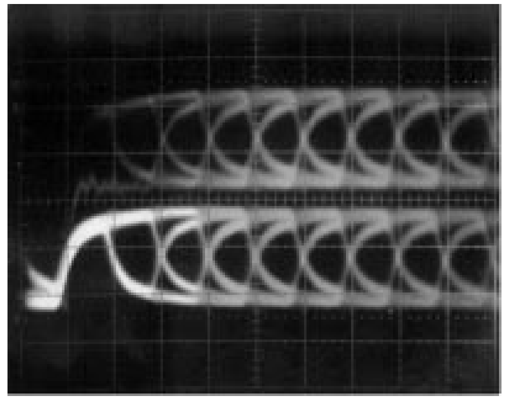

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FIGURE 10a. 50 MHz Signal at Receiver Input. (Intensification Due to Overlapping 10 MHz Signal) 20 ns/div. Horiz.

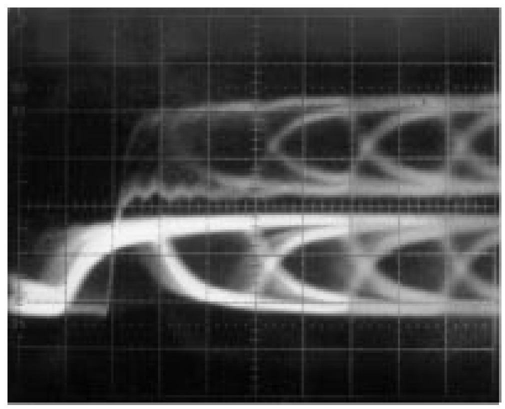

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FIGURE 10b. 50 MHz Signal at Receiver Input. 10 ns/div. Horiz. Peak-to-Peak Jitter is Less Than 40% Total.

Figures 10a and *10b* show the 50 MHz signal at the receiver input. The peak-to-peak jitter from all sources is about 40% maximum. *Figure 11a* shows the 50 MHz signal at the line input (differential mode) and *Figure 11b* shows the recovered 50 MHz signal at the receiver output. The effective jitter can be seen in the recovered signal. It is important to note that the F100250 exhibits no threshold shift which would contribute to bias distortion.

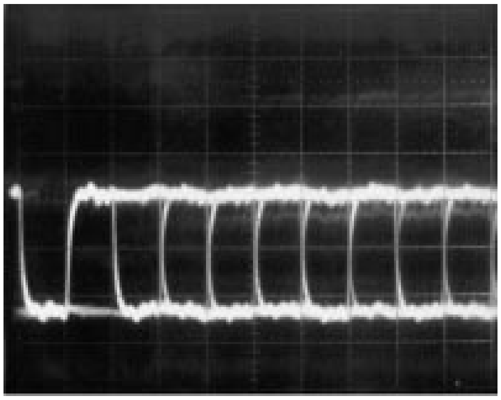

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FIGURE 11a. 50 MHz Signal at Transmitter Output (Line Input)

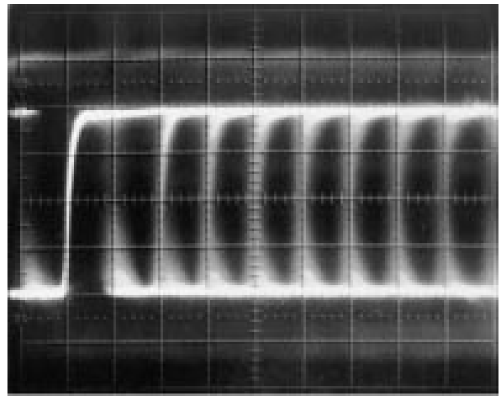

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FIGURE 11b. 50 MHz Signal at Receiver Output (Recovered Signal). Shows Effect of Jitter on Output.

APPENDIX A—TWISTED PAIR CABLES

Cable Type	Manufacturer	AWG Wire Size	Z_0 Ω	V_P	C_0 pF/ft	Attenuation 6 dBV Limit k Feet
TWISTED PAIR						
Unshielded	Com'l	28	100	0.66	15.5	0.77
		28	120	0.78	11.0	0.9
		26	106	0.66	12.0	2.7
		24	100	0.66	15.5	2.1
		24	120	0.66	12.8	2.5
		24	100	0.78	12.5	2.1
		22	100	0.66	15.5	3.0
Individually Shielded	Com'l	24	100	0.78	12.5	2.1
		25	124	0.66	12.2	1.9

APPENDIX B—COAXIAL CABLES

Cable Type	Manufacturer	Part Number	Z_0 Ω	V_P	C_0 pF/ft	Attenuation dB/100 ft @ 100 MHz
COAXIAL						
Single	Com'l	RG-59/U	75	0.78	17.3	3.0
	Belden	8281	75	0.66	21.0	2.7
Dual	Belden	9555	75	0.66	20.5	3.4
	Alpha	9845	75	0.66	20.5	3.4
Single	Com'l	RG-62/U RG-62A/U RG-62B/U MIL-C-17F	93	0.84	13.5	2.7
	Belden	9393	93	0.78	14.0	8.8
	Alpha	9063B	125	0.84	9.6	1.5
TWINAXIAL						
	Com'l	RG-22B/U	95	0.66	16.0	3.0
	Belden	8227 9207 9815	100	0.66	15.5	4.1
	Belden	9271 9860	124	0.66	12.2	5.0
	Belden	9182	150	0.78	8.8	4.3
	Belden	9851	200	0.76	6.7	4.0 @ 50 MHz
RIBBON						
	Belden	9K750XX	75	0.78	17.1	7.5
		9K930XX	93	0.78	14 ±2	5.0

APPENDIX C— MEASURING CABLE IMPEDANCE

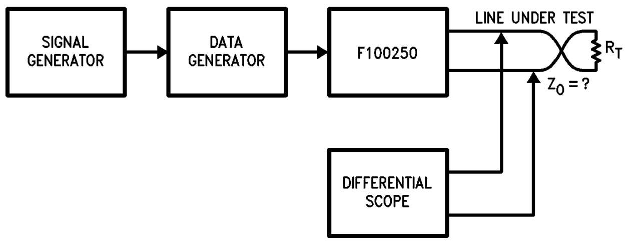
The impedance of a length of cable may be determined from simple measurements using the eye pattern. Either of two methods may be used. In one method the voltage reflection from a known termination is used to calculate Z_0. The second method uses direct resistance measurement to find Z_0.

Voltage (Indirect) Method

In the voltage method *(Figure C1)*, the signal generator frequency is set such that the unit interval of the eye pattern is about twice the round-trip delay of the line to be measured. The peak voltage (V_{peak}) of the eye pattern cell is measured *(Figure C2a)* with a known value termination resistor at the far end of the line. Next, the voltage at the end of the bit cell (V_{nom}) is measured. The line impedance can then be calculated using the following formula to about 5% accuracy.

$$Z_0 = R_t * ((2 * V_{peak}/V_{nom}) - 1)$$

For the waveform shown in *Figure C2b*, a 51Ω, 5% resistor is used as the termination. The peak line voltage is 390 mV. The voltage at the end of the bit cell is 240 mV. This gives a line impedance of 114.7Ω.



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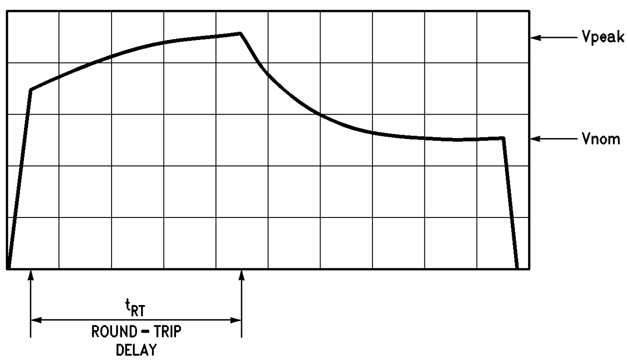
R_T = Suitable fixed resistor (voltage method) or 250Ω to 300Ω variable resistor (cermet potentiometer) for direct method

FIGURE C1. Line Impedance Test Setup



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$$Z_0 = R_t \left(\frac{2\,V_{peak}}{V_{nom}} - 1\right)$$

FIGURE C2a

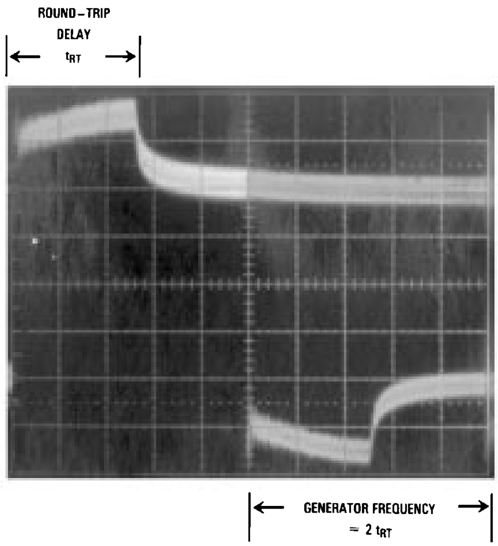

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FIGURE C2b. Voltage Method Waveform

Incidentally, the round-trip delay of the line, represented by the rising portion of the waveform, is 520 ns. Since the cable sample is 50 meters (164 ft.) long, the delay per foot can be estimated to be 1.6 ns per foot. (The data sheet for the cable sample used specified the delay of the pair as 1.575 ns to 1.6 ns per foot and the impedance as 101.5Ω to 105Ω.)

Resistance (Direct) Method

The same test setup is used as for the voltage method. A variable resistor is substituted for the fixed termination. The value chosen should be in the range of 250Ω to 300Ω.

$R_T < Z_0$

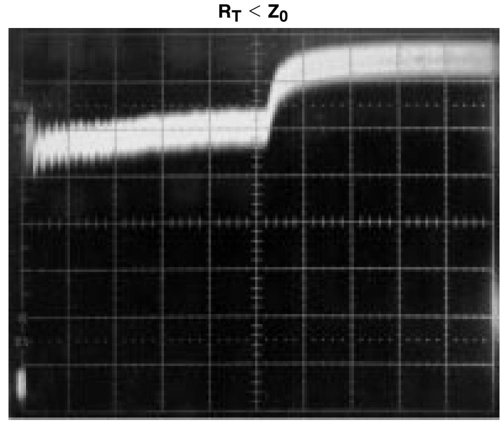

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FIGURE C3a. Under-Terminated

$R_T > Z_0$

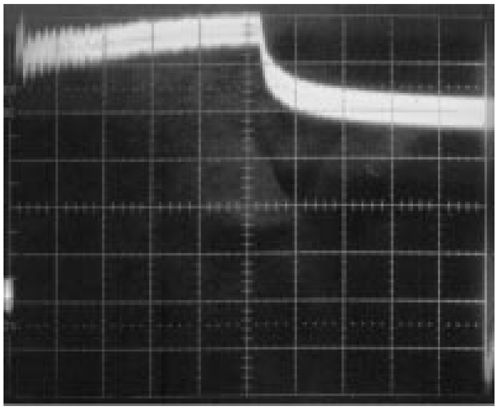

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FIGURE C3b. Over-Terminated

$R_T = Z_0$

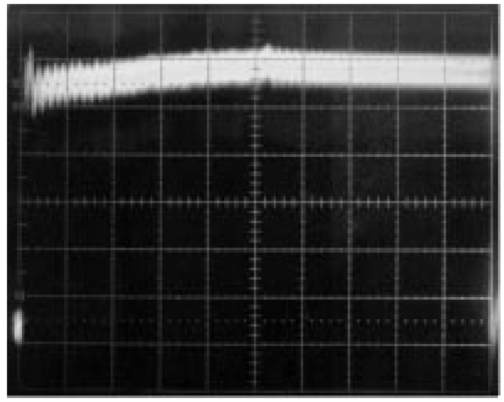

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FIGURE C3c. Terminated

Figures C3a through *C3c* illustrate the three possible conditions of termination which can be achieved. *Figure C3a* shows the under-terminated condition where the termination value is greater than the line impedance. *Figure C3b* is the over-terminated condition where the termination value is less than the line impedance. Finally, *Figure C3c* shows the condition where the termination is adjusted to match the line impedance. When this condition is achieved, the value of the termination variable resistor, and hence the line impedance, may be read using an Ohmmeter.

Using the direct method to measure the same pair sample gave a value of 105.9Ω for the line impedance. This method is preferred since it is simple and accurate.

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