FEATURE

INTERCONNECTION SYSTEM APPROACHES FOR MINIMIZING DATA TRANSMISSION PROBLEMS

Rf circuit design techniques reduce signal degradation and crosstalk to permit increased bandwidth in high performance digital systems

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Advances in digital logic and complex components, in conjunction with the extraordinary speeds of charge-coupled devices, emitter-coupled logic, and gallium arsenide technologies, provide the high performance demanded by today's computing environment. As the speed of logic families increases, signal rise and fall times and propagation delays decrease. The problem of getting data from one place to another in a high performance system becomes more difficult with this delay and rise time decrease, often as a direct result of the bandwidth of the interconnection system required to carry data pulses.

The graph in Fig 1 plots the trend of common logic families, extending to the latest in high speed technology. Lines marked N = 1 and N = 10 plot the equation of the bandwidth required to pass the pulse signal.

$$BW = 0.35N/t_r$$

In the equation, t_r equals the device rise time and N is the highest harmonic of the propagation frequency to be passed.[1] At N = 1, the system distorts the pulse rise and fall times, and barely passes the leading and trailing edges of the desired pulse. In the normal high performance situation, at N = 10, the system passes signals

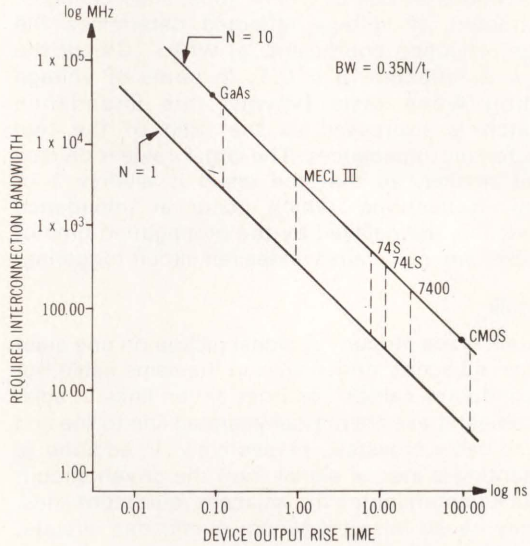


Fig 1 Fast logic demands high bandwidth. Graph plots approximate system interconnection bandwidth for maximum performance as function of device rise time for popular and more recent logic families. At N = 1 harmonic, system barely passes pulse rise time. At N = 10 harmonic, system passes rise time almost undisturbed

Transmission Line Factors

Characteristic Impedance

Characteristic impedance (Z_0) is a function of the geometry and the materials of a system. In a transmission line, two wires are surrounded by a dielectric material. Between the two wires is a capacitance, which is a function of the spacing between the wires, the dielectric constant of the material between them, and the effective plate area of the wires. In simple instances, plate area is equal to the diameter times the length of the wire, and capacitance is expressed as microfarads per unit length. The two wires have a mutual inductance that is essentially a function of the distance between the wires, their diameter, and the wire length (the magnetic permeability of the surrounding material is usually unity). This inductance could be expressed as microhenries per unit length. If a 1-unit section of wire had a capacitance of x microfarads, and an inductance of y microhenries, the impedance (Z_0) would be $Z_0 = (y/x)^{1/2} = A\,\Omega$. For a 2-unit section, the capacitance would be 2x and the inductance 2y; impedance would remain the same. Therefore, the impedance of a matched line is independent of its length. It could be said that each unit area of line cross section has its own characteristic impedance.

Reflection Coefficient

The reflection coefficient is a measure of how well the transmission line has achieved impedance control. A well-designed system, in which the circuits at both ends of a line are matched to the impedance of the line itself, provides for maximum energy transfer between the circuits. Furthermore (and more important for digital applications), a well-matched system prevents the generation of noise on the line, indicated by reflections at points where impedance changes. The amount of voltage reflected determines the voltage reflection coefficient, ϱ; when 10% of the voltage is reflected, $\varrho = 0.1$. In terms of voltage standing wave ratio (VSWR), this impedance mismatch is expressed as the ratio of the two characteristic impedances. The larger value is divided by the smaller, so that the result is always 1 or greater. Reflections, which occur at impedance mismatches, are delayed by the propagation time of the cable, and may cause undesired circuit triggering.

Crosstalk

Crosstalk is the amount of signal pickup on one quiet line from adjacent, driven lines in the same cable (for multiconductor cables), or from driven lines in adjacent cables; these are normally termed line to line and cable to cable crosstalk, respectively. In addition to representing a loss of signal from the driven circuit, crosstalk contaminates the adjacent, quiescent lines, and may cause false triggering, overloaded circuits, and interference. Crosstalk can be measured in both forward and backward, or reverse, directions. Backward crosstalk is measured at the input end of a quiet line with respect to a driving device. Here, coupling of the driven signal produces an output on the quiet line, which has the same polarity as the applied voltage, but is reduced in magnitude by a factor related to k_B, the back crosstalk coefficient.

Forward crosstalk, on the other hand, is related to the velocity of different modes of the propagating wave. In a homogeneous dielectric medium, the mode velocities are equal; consequently, forward crosstalk is zero. However, in an open transmission line system with composite dielectrics of both air and substrate, the forward crosstalk becomes a function of the transmission line length, the voltage rise time of the applied signal, and, indirectly, the circuit coupling. Although peak forward crosstalk can be fairly high, its duration is very short, (often less than t_r), so that little energy is actually transferred, and many circuits are unaffected.

Attenuation

Attenuation of digital pulses results in degradation or distortion of the pulse in terms of a loss in peak voltage, a slower pulse rise, or both. There are two possible sources for these problems: conduction loss in the dielectric and resistive loss in the conductor. Both are functions of frequency; dielectric attenuation varies directly, whereas conductor loss (skin effect) varies as the square root of frequency. The sum of these two loss parameters represents the total loss of a line section and is usually measured in dB per unit length. At frequencies of 1 GHz, the dielectric loss represents about one-third of the total attenuation. At 500 MHz, dielectric losses are down to one-fourth of the total. Thus, resistive losses tend to dominate below 1 GHz. With respect to conductor size, attenuation and its dc relatives, resistance and current carrying capacity, are essentially functions of wire size. As signal frequencies get higher, however, a greater portion of signal current is carried on the surface of the conductor (skin effect). A flat conductor, with its greater surface to volume ratio, thus provides lower high frequency attenuation than does a round wire of equal size.

Propagation Delay

Propagation delay is the time required for a pulse to travel through a transmission line system. It is the reciprocal of signal propagation velocity, and is, essentially, a function of the dielectric constant of the insulating material. Low-delay signal transmission is essential for high speed computer systems. In open transmission systems, where the insulation is not homogeneous, effective dielectric constant is an average value of the different materials. In most flat cables, some of the electromagnetic field escapes the insulation, and propagates in the air around the cable. This tends to decrease the dielectric constant, which increases the propagation velocity, but also increases the ground plane susceptibility, which may have the opposite effect. At high frequencies, when the dielectric constant of the insulation varies with the frequency, the different frequency components see different dielectric constants, and different propagation velocities cause pulse distortion.

easily, with almost no pulse edge distortion. For example, to use 74LS parts to maximum advantage, the system bandwidth must be extended from the very high frequency (vhf) range to the ultrahigh frequency (uhf) range mostly encountered by radio frequency (rf) circuit designers (and often unfamiliar to digital or data system hardware designers). As system speeds increase, system designers must familiarize themselves with parallel data transmission systems whose bandwidths easily extend into the microwave region.

Phase delay (time skewing) and ringing have a critical impact on system operation. In addition, parallel data transmission paths can produce crosstalk, significant noise in a "quiet" line induced from the active lines. Data integrity is maintained in typical systems by restricting ringing and crosstalk to about 10% of normal signal levels (−20 dB). High performance systems generally operate with relatively short runs in which data repeating amplifiers are not used along the basic transmission path and high noise margins are desirable. Amplitude losses in many systems, therefore, must be held to 20% to 25% (2 to 2.5 dB) along the transmission path.

When overall ringing, crosstalk, and signal levels cause signal degradation of more than about 30% at the chip receiving the data, noise margins become unacceptably high, and produce excessive error rates. High amplitude, wide bandwidth, and electrical noise environments must also be taken into account, but fortunately most system designs that eliminate crosstalk problems can eliminate outside electrical noise problems as well.

Failure to plan for the analog requirements of rf transmission at high frequencies when designing the interconnection system can create a system level malfunction that is almost impossible to cure through a retrofit. Often transmission problems do not show up until late in the design cycle, during board integration and test. Retrofit at that point would require significant changes in chip fanout and board layout densities; the selection and use of proper connectors adapted to wide bandwidths; and the use of special, low loss cable.

Data Transmission Problems

Circuit and systems designers unfamiliar with very high speed data transmission did not need to consider the effects of basic interconnection paths in previous designs; data transmission line problems do not exist over a signal transmission distance that is a very small fraction of the signal wavelength. However, system performance can be impaired as the wavelength of the highest frequency being transferred approaches the transmission path length. After the data transmission path exceeds a significant fraction of a wavelength, a transmission line approach must be introduced, with its attendant factors including characteristic impedance, attenuation, and reflection coefficient.

Fig 2 represents some critical signal path distances, and plots critical wavelength as a function of system cutoff frequency; the range of critical distances depicted extends from 8 ft (2 m) for 7400 logic, to 0.65 in (17 mm) for MECL III. The curve plots $\lambda_{1/8} = V/8f$, where V is the propagation velocity, and f is the rise time frequency being propagated. (Compare Fig 1.) The range of electrically equivalent distances for data transmission at each frequency extends from 23 ft (7 m) to 0.3 in (8 mm), as f varies from 3.5 MHz to 3.5 GHz.

Point A in Fig 2 represents a critical distance of 8 ft (2 m) at the 10-MHz frequency that represents the approximate lower end of useful bandwidth for a high performance 7400 transistor-transistor logic (TTL) system. Point B similarly represents the critical one-eighth wavelength distance of 2.3 ft (0.70 m) at 35 MHz, the approximate lower end of useful bandwidth in a high performance 74LS logic application. High performance 74S logic has a similar critical distance of 2.8 in (71 mm) for a transmission interconnection bandwidth of 350 MHz, shown at point C. Point D represents the high performance end of operation of MECL III logic, having a critical distance of 0.65 in (17 mm) at a frequency of 1.5 GHz. At these speeds, even the distance a signal travels through a connector is important.

Stray capacitance between signal conductors increases signal interference from one data line to another (crosstalk). As an example, 1.0 pF of stray capacitance between two conductors has a reactance of 450 Ω at 350 MHz and 45 Ω at 3.5 GHz. To illustrate, a 1.25-in (31.8-mm) length of typical ribbon cable, which has an approximate capacitance of 1 to 2 pF, can cause significant signal coupling from one parallel data line to another, and generate noise on the second line. Depending on the circuit impedance in the system, such signals can significantly reduce noise margins, thereby increasing error rate.

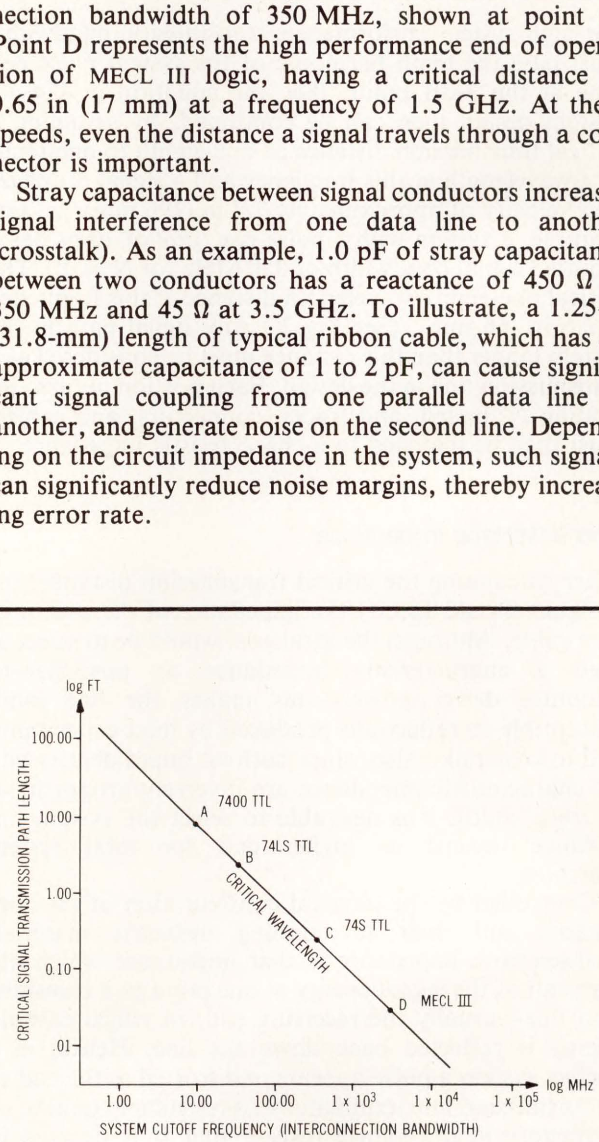


Fig 2 High bandwidth reduces critical path length. Graph plots critical signal path length as function of interconnection bandwidth, assuming critical path of one-eighth wavelength. Assumed system wave propagation velocity is 6.55 x 10⁸ ft/s, which is typical propagation velocity in transmission system

Design Recommendations

Attention must be paid to every detail in the rf transmission path, but only a few basic concepts are essential for adequate system design. Choosing characteristic impedance, the most important design decision, involves a small set of tradeoffs among the logic family used, printed circuit (PC) boards that are practical to fabricate, and the impedance of available cable/connector combinations. Determining the design approach for the required system transmission bandwidth involves a tradeoff between the performance dictated by the PC technology and an acceptable error rate and cost.

Transmission Bandwidth

For high system performance, transmission bandwidth must pass the tenth harmonic of the system pulse rise time at the 3-dB point. (See the equation in Fig 1.) System design then can be continued, by assuming a critical transmission distance of one-eighth to one-tenth of a wavelength at this frequency and a signal propagation velocity of approximately 7.7 in (196 mm)/ns. For example, a system with a pulse rise time of 10 ns has a required signal bandwidth of 350 MHz for N = 10. The critical one-eighth wavelength distance at this frequency is 2.6 in (66 mm). (See Fig 2.) Any signal path in the system longer than this distance must be considered as a transmission line in the design. Each portion in turn, including PC board, and traces, connectors, and cables, must then be matched to a characteristic impedance.

Characteristic Impedance

After calculating the critical transmission distance, the designer should specify the impedance of the transmission cable. Although the ideal goal would be to select as high a characteristic impedance as possible—to minimize driver power—this makes the line more susceptible to reflections produced by load capacitance and to crosstalk. Also, since both PC board density and PC characteristic impedance are inversely proportional to trace width, it is desirable to select the system impedance toward its higher end for total system matching.

Controlled by the physical configuration of the conductors and their surrounding dielectric material, characteristic impedance is that impedance which absorbs all of the signal energy at one point in a transmission line—usually the receiving end, in which case no energy is reflected back down the line. Hence, in a perfect system a pulse appears undistorted at the end of a transmission line terminated in a resistance equal to its characteristic impedance (except that it is delayed in time). In a real system with fast rise times, the receiving integrated circuit (IC) input capacitance prevents this ideal resistive match.

When the line is not matched, energy is reflected from the end of the line and appears to reflect up and down the line, with a delay between each reflection that is proportional to the length of the line; reflections appear as ringing (or even distinct steps) at the receiving end. In short, the distorted signal appears as a coherent noise on the signal line; when sufficiently high in amplitude, this noise can cause data errors.

The reflection coefficient ratio determines how well a transmission line is matched to its characteristic impedance. Practical high performance systems must use a specific characteristic impedance throughout the transmission path. As the signal passes from the sending IC through the PC board; connector, backplane, or cables; a second connector; the receiving PC board; and, finally, the receiving IC, it goes through a number of physical elements [Fig 3(a)]. Any significant section of the path must maintain the selected transmission characteristic impedance. An abrupt change in impedance is called a discontinuity. Energy is reflected from the point of each discontinuity, and creates an effect similar to that of an unmatched line.

Testing

Testing is best accomplished using the time domain reflectometry (TDR) technique. TDR applies a step pulse (generated from a known source impedance and usually less than 150 ps) to the transmission system. A wide bandwidth oscilloscope measures the resulting signal waveform. Since the pulse step is very rapid, it is possible to locate the position of the system impedance discontinuities and measure their reflection coefficients by observing the resulting signal. The step signal propagates down the line at a controlled velocity, and the reflections from any impedance discontinuity appear as reflected voltage at a time measurable from the start of the pulse. TDR can also observe inductive and capacitive loading effects, and is used universally to determine the quality of transmission possible in a given system.[2] This includes determining the degree of mismatch in PC board traces, connectors, and cables. Using TDR, the designer can verify how well the design has achieved a matched condition.

TDR, however, has limitations. One is that the system bandwidth of TDR limits the shortest length of an observable discontinuity. Another limitation is that TDR equipment may have a rise time much faster than that of the system under investigation, and consequently give too much detail concerning discontinuities that will have little effect on the actual system. This latter problem can be reduced by adding controlled filters to slow the TDR rise time and limit the range of output frequencies from the TDR step generator. TDR can also be used to study the system crosstalk effects. In this case, the step is placed on the active line, and its effects are measured on the quiet line.

Basic System Elements

Once the characteristic impedance is selected, in practice somewhere between 50 and 150 Ω, the cables, connectors, and PC board can be chosen. Only a few approaches to PC boards are practical, and cable and connector choices are often interdependent (especially in

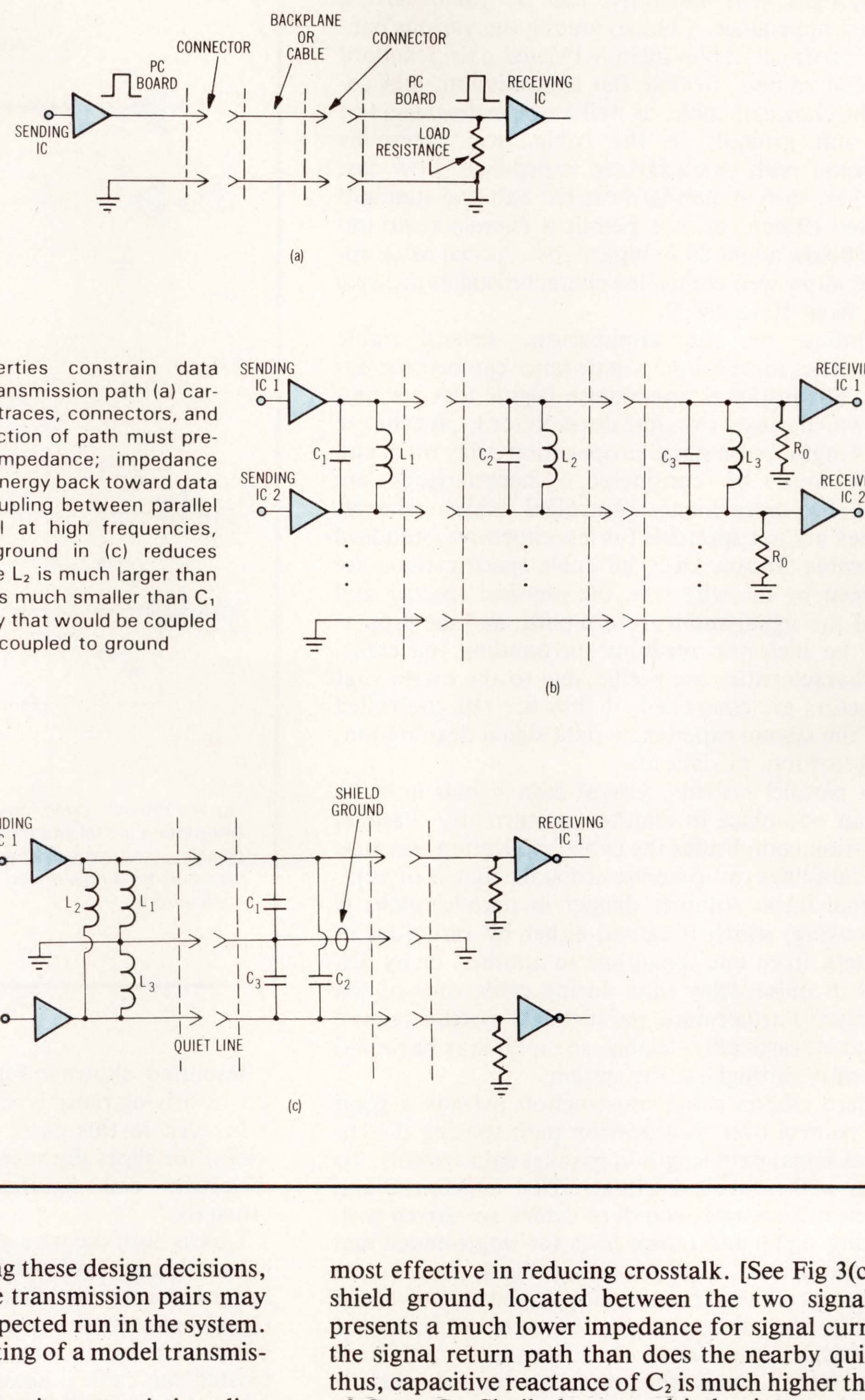


Fig 3 Signal path properties constrain data transmission. Typical data transmission path (a) carries signal through PC board traces, connectors, and backplane or cable. Each section of path must present same characteristic impedance; impedance discontinuity reflects signal energy back toward data source. In (b), capacitive coupling between parallel lines becomes more critical at high frequencies, causing crosstalk. Shield ground in (c) reduces crosstalk because inductance L_2 is much larger than L_1 or L_3 and capacitance C_2 is much smaller than C_1 or C_3; therefore, signal energy that would be coupled from path to path is instead coupled to ground

parallel data systems). Following these design decisions, PC board, connector, and cable transmission pairs may be configured for the longest expected run in the system. Critical to this process is the testing of a model transmission system.

Several other important circuit transmission line characteristics to consider are PC board microstrip (and other) rf techniques; specialized data about rf characteristics of standard connectors or special connectors designed for high speed use; and specialized rf characteristic data about standard cables such as ribbon or flat conductor types, or special cables designed for high speed data use. Also, crosstalk at high frequencies can be reduced by spacing the parallel data conductors farther apart, reducing the coupling in either PC board traces or cables [Fig 3(b)]. Shielding bypasses the mutual signal line capacitance to signal ground and is most effective in reducing crosstalk. [See Fig 3(c).] The shield ground, located between the two signal lines, presents a much lower impedance for signal currents to the signal return path than does the nearby quiet line; thus, capacitive reactance of C_2 is much higher than that of C_1 or C_3. Similarly, mutual inductive reactances L_1 and L_3, created by the effects of currents in either signal line and in the ground line, will be much smaller than the inductive reactance between the signal lines, L_2. An ideal transmission line configuration would be a totally shielded or coaxial cable, because the surrounding shield would reduce intercoupling dramatically.

Cables

Cables carry data between widely spaced sections of a subsystem; therefore, they are likely to be the most critical portion of the transmission path. In high performance systems, it is imperative that the cables have a controlled impedance. Choices among the various controlled impedance cables include twisted pair, standard and special ribbon, flexible flat transmission, and coaxial. The choice of cable, as well as the arrangement of signals and grounds in the cable, will determine transmission path characteristic impedance. Low cost approaches, such as standard twisted pair and standard unshielded ribbon, do not permit a characteristic impedance below about 80 Ω; higher cost, special cable approaches allow well-controlled characteristic impedance ranging from 50 to 150 Ω.

Depending on the application, several cable characteristics in addition to impedance can become important. For instance, attenuation (signal loss per unit length, which causes rise time degradation), capacitance per unit length, pulse signal propagation delay time, and crosstalk must be considered. Characteristics are specified for more highly controlled cables, whereas often they are not specified for less expensive, standard ribbon cable. In any case, all cable specifications are determined by the wire size, the physical spacing and shape of the signal and its return path, and the properties of the dielectric medium surrounding the cable. Cable characteristics are predictable to the extent that these factors are controlled; if they are not controlled closely, the system experiences data signal degradation, signal distortion, or data loss.

In bit parallel systems, several data signals must be sent from one place to another concurrently. Parallel transmission complicates the cable application, because several data lines can generate crosstalk signals in adjacent signal lines. Another danger in parallel paths is pulse skewing, which is caused either by variations in path length from one signal line to another, or by differences in pulse delay time during cable runs of the same length. Furthermore, parallel data systems require cable that is reasonably flexible, so that it may be routed conveniently throughout the system.

Standard ribbon cable construction permits a good deal of control over transmission path spacing dimensions and signal path length in parallel data systems. To maintain well-controlled characteristic impedance and low levels of crosstalk, standard cables are driven with alternating signal and return lines for single-ended and differential drivers, as illustrated in Fig 4. Such an arrangement produces characteristic impedance of about 85 Ω. Fig 5(a) illustrates the expected variation of characteristic impedance for standard polyvinyl chloride (PVC) ribbon cable, and also for polyethelene cable clad with PVC. The wire diameter, d, varies from 0.02 in (0.5 mm) for #24 gauge to 0.01 in (0.3 mm) for #30 gauge wires.

Crosstalk is one of the major limitations of ribbon cables. Even when carrying relatively slow rise time pulses, such as 10 ns for 20-ft (6-m) distances, the crosstalk can be higher than 30% for standard cables. Fig 5(b) indicates typical forward crosstalk (at the far end) of a ribbon cable driven and loaded by its characteristic impedance.

Typical cable also has limited bandwidth capability, as illustrated by the pulse degradation with the distance

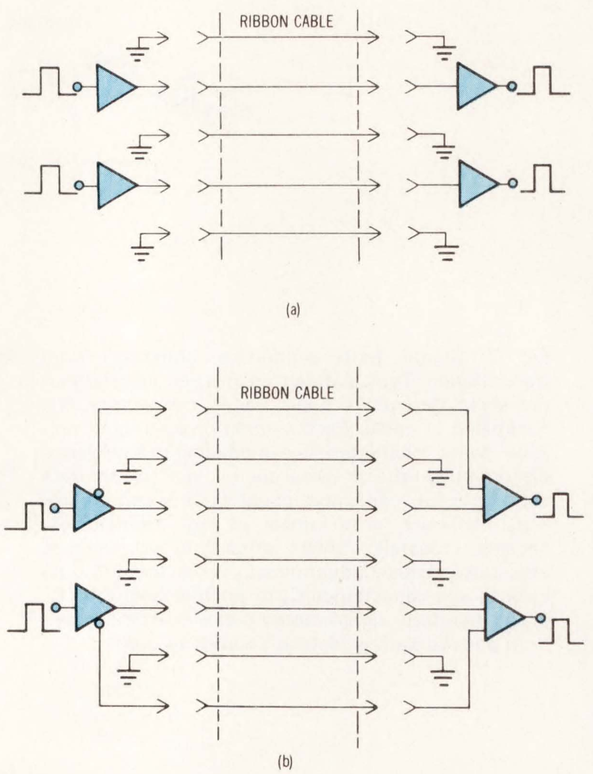
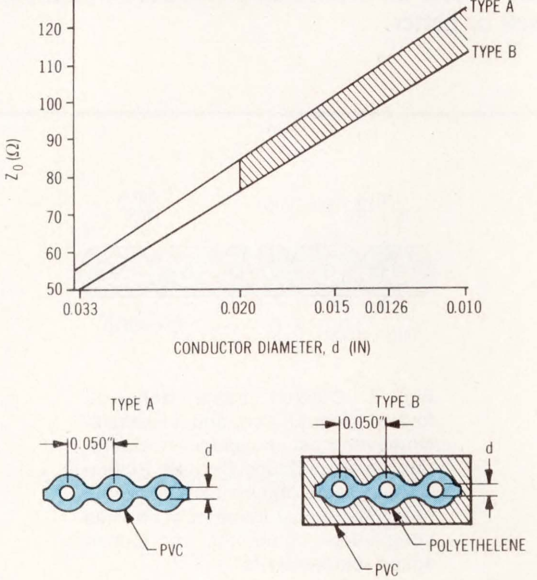


Fig 4 Ribbon cable controls path spacing and length. Alternating signal and ground return lines give ribbon cable closely constrained characteristic impedance and low crosstalk for single-ended drive (a) and differential drive (b) applications

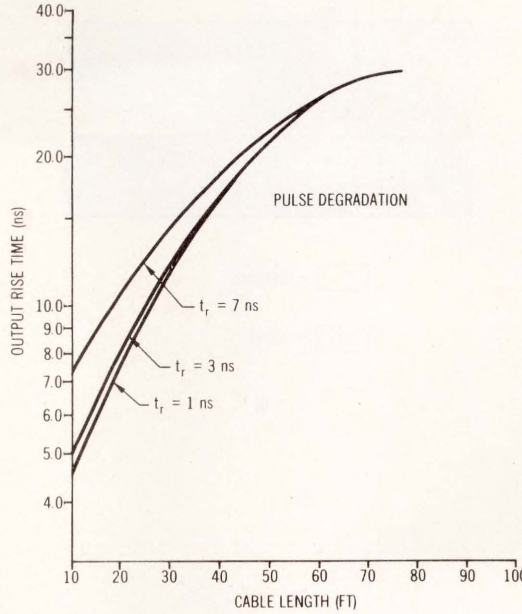
transmitted, shown in Fig 5(c). Note that the rise time of a 1-ns driving pulse is severely degraded at 10 ft (3 m). Yet, even in this case, standard ribbon cable can be useful for short distances. Flat ribbon cables with dual dielectrics can significantly reduce the flat cable crosstalk.[3]

Unless both the wire and the twisting are subjected to close tolerance, twisted cable has the disadvantage of a lower control on its characteristic impedance and actual signal path length variations that introduce the possibility of pulse skewing. Characteristic impedance for twisted pair cable is about 100 to 120 Ω. These cables offer excellent shielding characteristics from low frequency external fields, but are little better than cabled wire pairs in their susceptibility to crosstalk. Although they are acceptable for single-signal applications, they are not recommended for parallel data transmission.

Special ribbon and flat, flexible cables meet requirements for transmission characteristics in high performance systems, which can be divided into two basic classes: those with flat wide conductors (sometimes including a parallel signal ground), and those with a surrounding shield for the signal conductor (coaxial cable). Cables with flat conductors are referred to as flexible

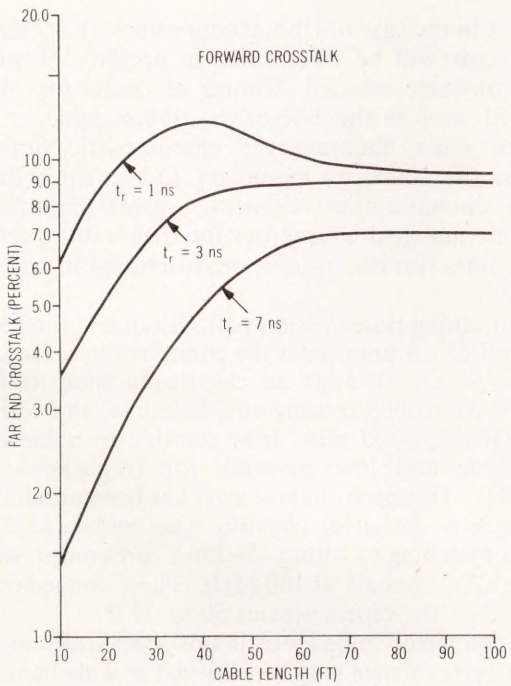


(a)



(c)

Fig 5 Ribbon cable properties. Graph in (a) plots characteristic impedance of unshielded ribbon cable as function of conductor size. PVC covering on clad ribbon cable increases dielectric constant and lowers characteristic impedance. Graph (b) plots ribbon cable crosstalk as function of cable length for 1-, 3-, and 7-ns rise times. In contrast, 20 ft of standard cable can have 30% crosstalk for rise times as slow as 10 ns. Graph (c) shows limited bandwidth of ribbon cable by plotting output rise time as function of cable length for 1-, 3-, and 7-ns input rise times. Here, in only 10 ft, 1-ns pulse suffers substantial degradation



(b)

transmission cables. They consist of etched copper on polyester film, and can have a parallel set of ground returns or a ground plane placed opposite them. Characteristic impedances of the etched copper cables range from 30 to 150 Ω. (See Fig 6.) Flat construction permits excellent control over the cable characteristics, and the square shape yields a lower attenuation factor. At high frequencies, the signal current tends to flow near the outside surface of the conductor (skin effect), and the square shape represents a surface area significantly larger than that of a round wire.

Coaxial cable performs even better than special ribbon and flat, flexible cable. Normally, the space required by individual connectors, in addition to the cost to assemble them, eliminates coaxial cable from consideration for parallel data systems, but ribbon coaxial cable with companion connectors creates a viable solution to these drawbacks. Manufactured with characteristic impedances of from 50 to 95Ω, cables are available with from 6 to 26 signal paths. A parallel coaxial cable with full shield coverage appears in cross section in Fig 7. When examining the characteristics of a 10-ft (3-m) section of 50- and 91-Ω cable, very low forward crosstalk (less than 0.1% and 0.8%, respectively), and only a small amount (400 ps) of rise time degradation, are evidenced at input pulse times of 1 ns. Separate coaxial cables on individual connectors would perform as well or better.

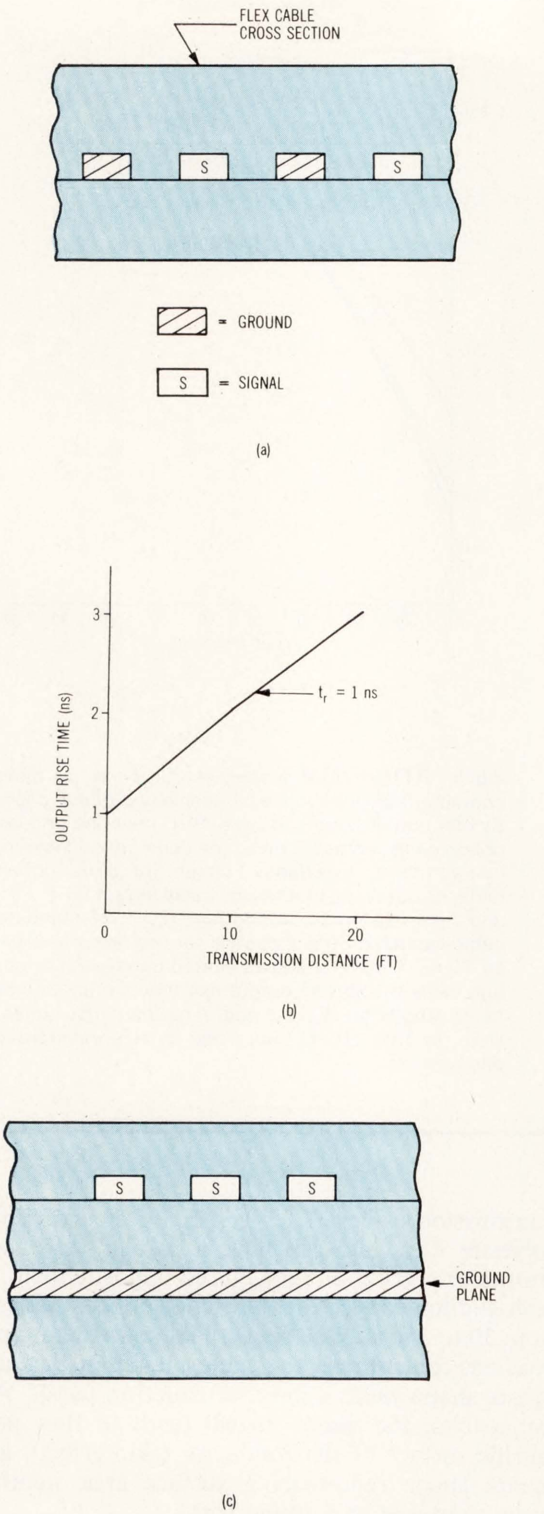


Fig 6 Flexible cable lowers degradation and crosstalk. Skin effect of square shape (a) reduces attenuation, and controlled rf properties increase bandwidth, causing less distortion of leading edges when high speed pulse travels relatively long distance (b). In (c), addition of ground plane reduces crosstalk to less than 4% for 10-ft cable carrying pulse with only 1-ns rise time

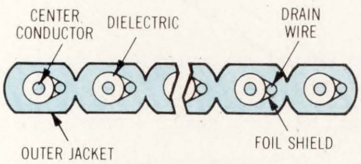


Fig 7 Coaxial cable achieves lowest degradation and crosstalk. However, cost and space considerations may preclude its use. Ribbon coaxial cable shown here performs nearly as well at lower cost and has companion connector to reduce space requirements

Connectors

Except in the case of ribbon connectors, the companion connector will be determined in accordance with the type of cable selected. Choice of connectors may be limited, even in the case of the ribbon cable, since the design must maintain the characteristic impedance across the connector boundary to prevent reflections from discontinuities. Generally, cable manufacturers provide matched connectors for their cables, and they sometimes furnish low cost, mass terminating capability as well.

Shortening transmission path distance is the best way to limit discontinuities at the connector interface. If the signal is sent through an electrically short connector with a controlled spacing and dielectric, and with nearby signal ground pins, it is possible to achieve good matching and low crosstalk for frequencies up to 10 GHz. This performance level has been obtained with a specially designed, chevron type connector that exhibits matching to within 4% for a 70-ps pulse, with less than 10% crosstalk at 100 MHz. These connectors have characteristic impedances of 50 or 75 Ω.

If connector impedance is not specifically matched, good performance may be achieved at wide bandwidths with the use of compensating pads on the PC board that receives the signal. At sufficiently low frequencies, the pins on the connector generally appear to be inductive, hence a capacitance pad on the PC board can add reactance to compensate for, or tune out, pin inductance using TDR method. Since a connector often represents a discontinuity in a transmission line, it is frequently measured using this technique to tune circuit performance. A connector can compensate for the transition from coaxial cable to a PC board; signal returns are placed adjacent to the signal connector pin and soldered to the ground plane via plated-through holes. Fig 8(a) shows the measured reflection coefficient (degree of match) of a connector system before compensation. The large peak indicates the inductance of the pin section. Notice that two input signal rise times were used in the measurement, 650 ps and 350 ps. Since the 350-ps trace contains more high frequency components, it shows a greater mismatch in the region of the connector and higher resolution. Fig 8(b) shows the results of the compensated system; the reflection coefficient approaches 0 (a perfect match) more closely, even for the 350-ps exciting pulse. At this point, the connector has achieved the capability of a matched bandwidth of at least 500 MHz.

PC Board

Microstrip transmission line technique is implemented on a 2-sided PC board by placing a ground plane on one side of the board, and a trace over it. By varying the trace width, and considering trace thickness, board thickness, and the dielectric constant of the board, the designer can create a practical transmission line. The approximate width of trace required to achieve a given characteristic impedance is illustrated in Fig 9, which shows trace requirements for the two standard board thicknesses of 0.06 in (2 mm) and 0.03 in (0.8 mm) on G-10 epoxy and 1-oz copper.[4] For 2-oz copper, the traces required are slightly narrower. Because the copper is not immersed in a constant dielectric, simple predicting equations for microstrip are relatively inaccurate. Often, the dielectric constant is not known beforehand; designers must experiment by building up simple transmission sections using known boards and traces, then fine tuning them by the TDR method for use in a critical application.

One interesting note is the rather large size of traces required for the thicker boards. For example, for a characteristic impedance of 50 Ω and a 0.06 in (2-mm) thick board, a 0.1-in (3-mm) trace is required. An 8-bit bus of such traces, separated for crosstalk, would occupy considerable board area. For this reason, and because of the physical constraints of other interconnection elements, designers tend to choose the highest target system characteristic impedance consistent with practical board fabrication techniques, crosstalk, and other interconnection limitations.

It is possible, by using at least a 3-layer board, to achieve a more predictable transmission line with the stripline approach (Fig 10). Striplining requires a ground plane on both sides of a conducting trace, thus immersing the trace in a constant dielectric, and making prediction easier. In the PC board, the trace lengths will determine the delay of the signal pulse; this effect can also become critical in high performance systems. For example, if the delay is about 150 ps/in, differences in the trace lengths of 1 in (25 mm) will produce skewing of 150 ps.

Thus, the designer must select a characteristic impedance for the board traces, make the traces as short as possible with sufficient separation to avoid crosstalk, and keep their lengths approximately the same for parallel signals. Stripline techniques, and even separate power distribution layers, will minimize power supply noise problems.

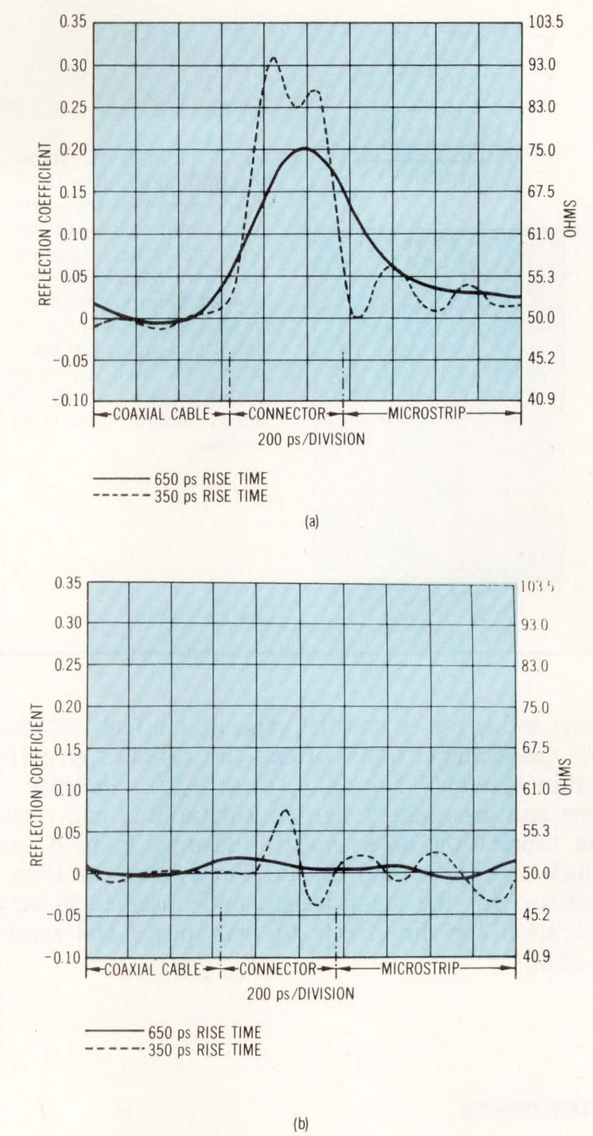
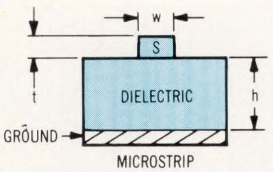


Fig 8 Connector compensates for impedance mismatch. Connector is used to compensate for mismatch at transition from coaxial cable to PC board. In (a), TDR plot before compensation, peaks resulting from connector pin inductance are shown. In (b), TDR plot of matched configuration shows effect of capacitance added to overcome pin inductance, achieving almost perfect match

Although this discussion assumes that all lines in the parallel data system are driven by steps from a driver of the characteristic impedance and a receiver loaded by its characteristic impedance, real drivers, in practice, have ringing and do not necessarily present the desired output impedance. Resistors used for matching can have either



Z_0 (Ω)	LINE WIDTH, w (THOUSANDTHS OF IN)	
	0.06" BOARD	0.03" BOARD
50	103	47
55	89	41
60	77	35
65	66	30
70	57	26
75	49	22
80	42	19
85	36	16
90	31	14
95	27	11
100	23	10

Fig 9 Microstrip transmission line approach. Trace thickness (t), width (w), board thickness (h), and dielectric constant determine microstrip properties; approximate characteristic impedance is listed as function of trace width for two common thicknesses of G-10 epoxy board with 1-oz copper

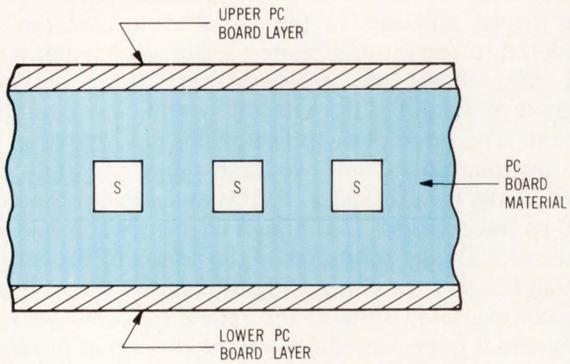


Fig 10 Stripline approach. Positioning ground plane on both sides of signal trace gives more predictable transmission line performance and better shielding, but at significantly higher cost because it requires multilayer board

series inductive or parallel capacitive effects. ICs have input capacitance that is often significant in high performance systems.[5] Normally, these effects increase rise time and crosstalk design considerations, and require the fabrication of a model transmission path upon which to base a final design. The system can then be matched for the ideal design, and observed in actual operation for the effects of real source and receiver loading.

Summary

Design plans for high performance, parallel data systems can compensate for the problems induced by increased transmission bandwidth. The process is lengthy and complex, and should be started early in the project. After characteristic impedance, attenuation, and crosstalk parameters have been traded off, and cables, connectors, and PC board layouts selected, a model transmission system should be established. The TDR technique will help the designer to obtain good, matched performance, and attain his final system design goal.

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