Using HOTLink™ with Long Copper Cables

Overview

The use of HOTLink™ data communications products to drive copper media is documented in a Cypress application note titled “Driving Copper Cables with HOTLink.” Long transmission lines (those that cannot be treated as lossless) present additional design concerns. The special characteristics and concerns of operation with long copper cables are covered here in this application note. This application note is also expected to be used in conjunction with a companion document titled “HOTLink Design Considerations.”

Primary Topics

The primary topics covered in this application note are

- Signal propagation
- Attenuation/Dispersion
- Equalization

Signal Propagation

Communication on short lengths of copper media allow the transmission line to be treated as lossless; i.e., a 1V square wave driven at one end of the cable comes out the other end with the same amplitude and waveshape. This is based on the simple relationship for transmission line impedance listed in *Equation 1*.

$$Z_O = \sqrt{\frac{L}{C}} \qquad \text{Eq. 1}$$

Real life transmission lines are not lossless. They contain numerous parasitic elements that cause a signal to distort as it propagates down the transmission line. When dealing with long cables, this equation must be modified to take into account the actual parasitics present in the transmission line. This places series-R and shunt-G components back in the calculation as shown in *Equation 2* (Reference 1).

$$Z_O = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \qquad \text{Eq. 2}$$

Loss Factors

This equation gets us bit bit closer to reality, but it assumes that the L, R, C, and G elements for a transmission line remain constant over frequency. In reality these “constants” often vary with frequency and are modified by four secondary loss factors:

- Skin effect
- Proximity effect
- Radiation loss effect
- Dielectric loss effect

Skin Effect

Skin effect is a current flow phenomenon where the cross-sectional current distribution in a conductor is affected by frequency. The higher the signal frequency, the higher the concentration of current on the surface of the conductor.

Skin effect is usually modeled as a dividing line that specifies the depth from the conductor surface where all current at a specific frequency is concentrated. In reality there is always some current flow in all parts of the conductor. At the higher frequencies most of it is concentrated at the surface.

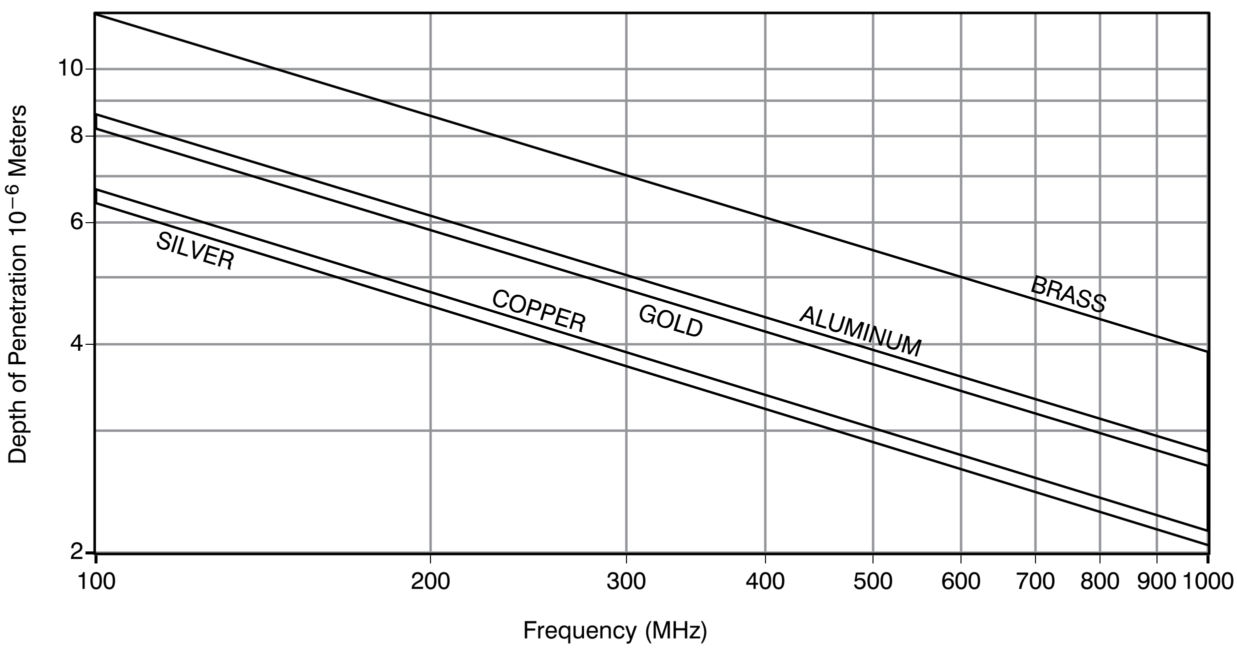

Figure 1. Effective Skin Depth

The effective skin depth is calculated using *Equation 3* (Reference 5).

$$d = \frac{1}{\sqrt{\pi f \mu \sigma}} \qquad \text{Eq. 3}$$

where:

μ = magnetic permeability of the conductor and

σ = conductivity of the conductor

Plotting effective skin depth over frequency (log/log scale) for a few common conductors (as shown in *Figure 1*) shows an interesting effect: all the lines are parallel. This is because the effective skin depth is directly proportional to the square root of frequency (Reference 2).

This change in the skin depth increases the conductors resistance as frequency is increased. This resistance change over frequency generates most of the attenuation losses in a cable (the L and C reactances are assumed to be lossless).

Figure 2 shows a frequency response plot of a few common cable types. The attenuation slope is approximately 0.5 for most of the cable types. This holds true for most standard sized cable constructions. For cables with composite plated conductors (like the RG179 cable) with various plating types (silver over copper over steel) the slope is modified by the changing current distribution in the different conductor types.

Proximity Effect

The proximity effect is caused by the current generated forces in adjacent conductors. Here the current distribution within a conductor is altered by the current present in a nearby conductor. This current redistribution works in conjunction with skin effect losses to further attenuate a signal. This loss factor does not effect coaxial cables but does effect twisted/parallel-pair cables, especially at higher frequencies. Generally the closer the conductors are and the higher the frequency, the greater the loss.

Radiation Loss Effect

Radiation loss is that signal lost due to electromagnetic radiation. This primarily effects unshielded-pair cables, or cables with poor shielding effectiveness. This loss type is often affected by those materials in close proximity to the transmission line.

For balanced transmission lines, it is also affected by the current balance within the two conductors in the transmission line. Any mismatch in amplitude or phase between the signals in the two conductors will

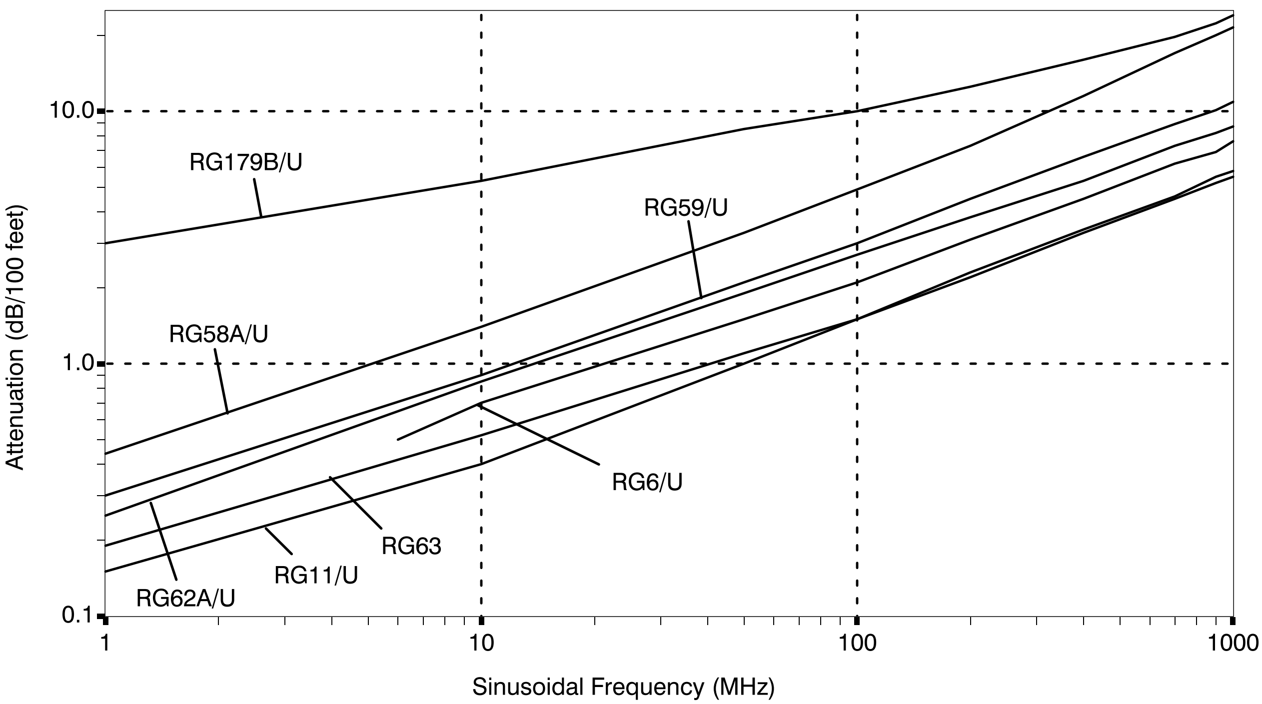

Figure 2. Coaxial Cable Attenuation Characteristics

radiate energy instead of propagating that energy down the transmission line.

Dielectric Loss Effect

Dielectric losses are those caused by the shunt conductance in the cable. This is represented by the G parameter in the impedance calculation in *Equation 2*. The loss mechanism here is current leakage through the dielectric. This loss is frequency sensitive and increases with frequency.

Reactance Factors

Just as the cable resistance and conductance vary with frequency, so do the inductance and capacitance. Both tend to decrease slightly with increasing frequency.

The change in inductance is due to the changes in skin effect, proximity effect, self inductance, and radiation loss. The change in capacitance is due to the dielectric constant of the dielectric spacer changing with frequency. The amount of capacitance change varies with the type of dielectric and the range of frequencies (Reference 1).

Signal Effects

These attenuation characteristics do more than just degrade the amplitude of a signal as it travels down a transmission line. They also affect the waveshape by distorting the rising and falling edges. The amount of the distortion is actually predictable, but it requires transformation of the source signal from the time domain to the frequency domain. This transformation is done using Fourier analysis.

Some of these effects may be illustrated using two simple square wave patterns. The first pattern is based on the highest frequency data pattern that can be sent, a continuous 0101 (D21.5 character) pattern. Using a 30-MHz byte-clock this pattern is equivalent to a 150-MHz square wave. The second pattern is based on the lowest frequency data pattern that can be sent, a continuous 0000011111 (K28.7) pattern (Reference 6). This pattern ends up being an exact match in period to the source clock (30 MHz) with a fixed 50% duty cycle.

Because the input waveforms are not true square waves, time constant curves based on a natural logarithm were used to synthesize the the rising and fal-

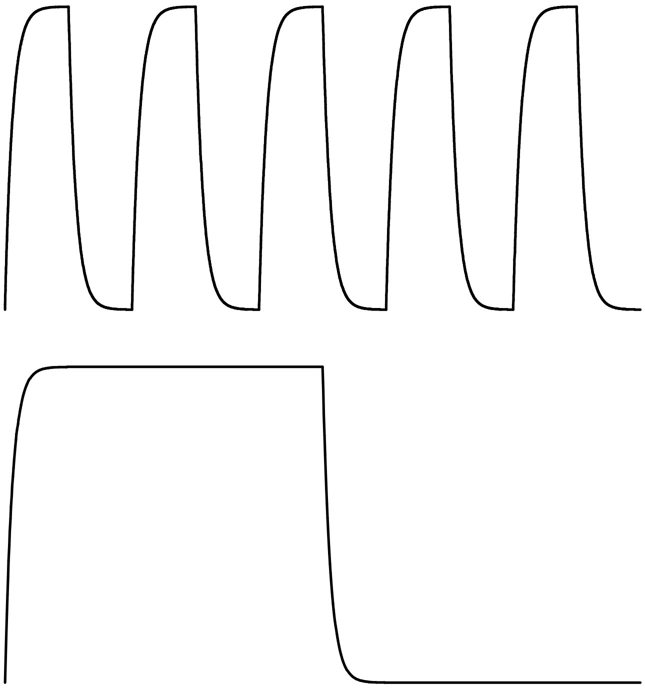

Figure 3. Synthesized D21.5 and K28.7 Waveforms

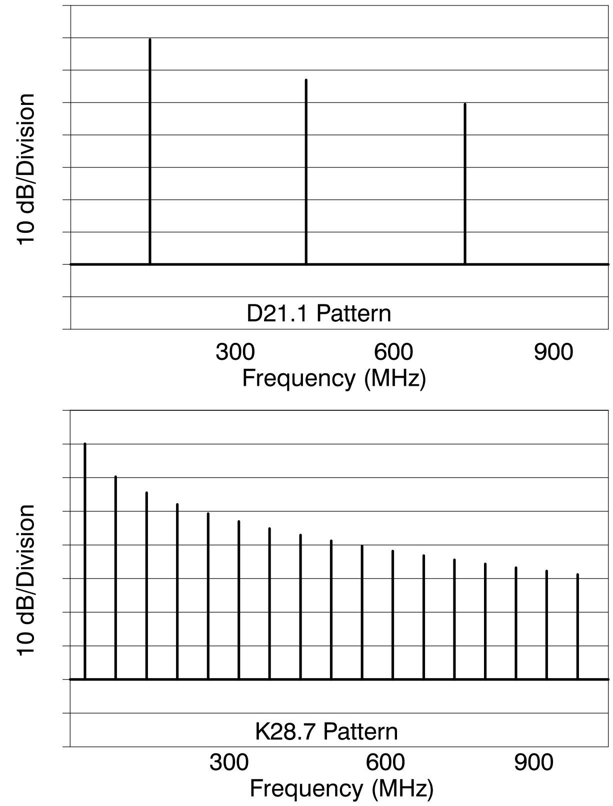

Figure 4. FFT Spectrum of Synthesized D21.5 and K28.7 Patterns

ling edges. These rising and falling edge equations are listed in *Equations 4* and 5 respectively.

$$T_R = 1 - e^{(-t/T)} \qquad \text{Eq. 4}$$

$$T_F = e^{(-t/T)} \qquad \text{Eq. 5}$$

In these equations, T represents the time constant for rise and fall time. For the waveforms generated for this example, a T of 400 ps was used. *Figure 3* illustrates the signals generated with these equations for both D21.5 and K28.7 characters (300-Mbit/second bit-rate).

Running a 4096 point FFT on these waveforms yields the spectral components in *Figure 4*. The vertical axis here is plotted on a log scale and shows the magnitude of the phasor at each spectral point.

Unlike a spectrum analyzer which only displays the magnitude of the spectral components, an FFT of a waveform yields both magnitude and phase in rectangular form as a complex number. To plot this information requires conversion to polar notation of magnitude and phase angle. This calculation of the magnitude portion is done using *Equation 6* (Reference 7).

$$Magnitude = \sqrt{\mathrm{Re}^2 + \mathrm{Im}^2} \qquad \text{Eq. 6}$$

An FFT is based on numeric analysis rather than a physical measurement and will calculate signal components with an amplitude of zero. Because Log(0) is equal to $-\infty$, a calculated FFT does not have a noise floor. To plot the results in a usable form requires the addition of an artificial noise floor to present the points of interest on a reasonable scale. To allow a better comparison with a real life environment, the noise floor in *Figure 4* is set at −80 dB.

Attenuation Effects

Now that the relative signal amplitude of each of the spectral components is known, a correction factor, based on the attenuation generated by a length of cable, can be applied to the spectral components. This attenuation is applied to the magnitude of the vector. A separate correction factor must be applied to the phase component.

Examination of a cable vendor's catalog will find a table for each cable listing attenuation at a few specific frequencies. The vendor's list of one such cable is found in *Table 1* (Reference 3). This information would be very helpful if the frequencies listed just happened to match up with the frequency components present in the signal being evaluated. Unfortunately this is rarely the case. Instead what must be done is to translate the table back into its transfer function, and use this function to calculate the attenuation at the specific frequencies of concern.

From *Figure 2* it is understood that that transfer function for a cable (in most cases) is approximated by a straight line, when plotted in log/log format. Geometry allows this line to be described in multiple ways, either by two points or as a slope and offset.

The manufacturer's attenuation data listed in *Table 1* is the same data that is plotted in *Figure 2*. Because this curve has few inflections, any of the points listed in the table may be used to approximate the transfer function. Since the data is plotted on a log/log scale, the calculations must be based on the log of both the frequency and the attenuation as shown in *Equation 7*. *Equation 8* calculates the slope for this cable type using data points at 10 MHz and 400 MHz (both at 100 meters).

Table 1. Attenuation for Belden 9659 Cable (RG59-type)

Frequency (MHz)	Nominal Attenuation	
	dB/100 Feet	dB/100 meters
1	0.3	1.0
10	0.9	3.0
50	2.1	6.9
100	3.0	9.8
200	4.5	14.8
400	6.6	21.7
700	8.9	29.2
900	10.1	33.1
1000	10.9	35.8

$$slope = \frac{Y_2 - Y_1}{X_2 - X_1} = \frac{\log(A_2) - \log(A_1)}{\log(F_2) - \log(F_1)} \qquad \text{Eq. 7}$$

$$\frac{1.3365 - 0.4771}{8.6021 - 7} = \frac{0.8593}{1.6021} = 0.5364 \qquad \text{Eq. 8}$$

The slope for most copper cables is around 0.5. (If only one attenuation data point is available, assuming 0.5 for a slope will get you close to the actual attenuation at other frequencies.)

With the slope available it is now possible to calculate the offset using *Equation 9*. The result as calculated at 400 MHz is shown in *Equation 10.*

$$offset = (log(F) \cdot slope) - log(A) \qquad \text{Eq. 9}$$

$$(8.6021 \times 0.5364) - 1.3365 = 3.278 \qquad \text{Eq. 10}$$

With the slope and offset now available, it is possible to calculate the attenuation per unit-distance at any frequency using *Equation 11*.

$$Attenuation(dB) = 10^{(log(Frequency) \times slope - offset)} \qquad \text{Eq. 11}$$

Note: Because all the previous calculations were based on 100 meter distances, the numbers generated here give the attenuation for 100 meters of RG59 cable at any frequency. These numbers may be scaled linearly to get the attenuation at any other length of cable.

The waveforms in *Figures 3* and *4* have symmetrical rise and fall times and therefore only contain odd harmonics. For the 30-MHz signal this yields harmonics at 30 MHz, 90 MHz, 150 MHz, 180 MHz, etc. The calculated attenuation for these harmonics (through 1 GHz) are listed in *Table 2*.

By applying these attenuation amounts to the specific signal components it is possible to determine the signal's spectrum at other points on the cable. These calculations were performed assuming a 100 meter length of cable to generate the spectrums shown in *Figure 5*.

By using an IFT (inverse Fourier transform) on these new spectrums it is possible to reconstitute the time domain form of the signal. If the same phase components are used with the attenuated amplitudes, the waveforms in *Figure 6* are generated (Reference 7).

Table 2. Calculated Attenuation for Belden 9659 Cable (RG59-type)

Frequency (MHz)	Nominal Attenuation	
	dB/100 Feet	dB/100 meters
30	1.64	5.40
90	2.96	9.75
150	3.90	12.8
210	4.67	15.4
270	5.34	17.6
330	5.95	19.6
390	6.51	21.4
450	7.03	23.1
510	7.51	24.7
570	7.98	26.2
630	8.42	27.7
690	8.84	29.1
750	9.24	30.4
810	9.63	31.7
870	10.0	32.9
930	10.4	34.1
990	10.7	35.3
1050	11.1	36.4

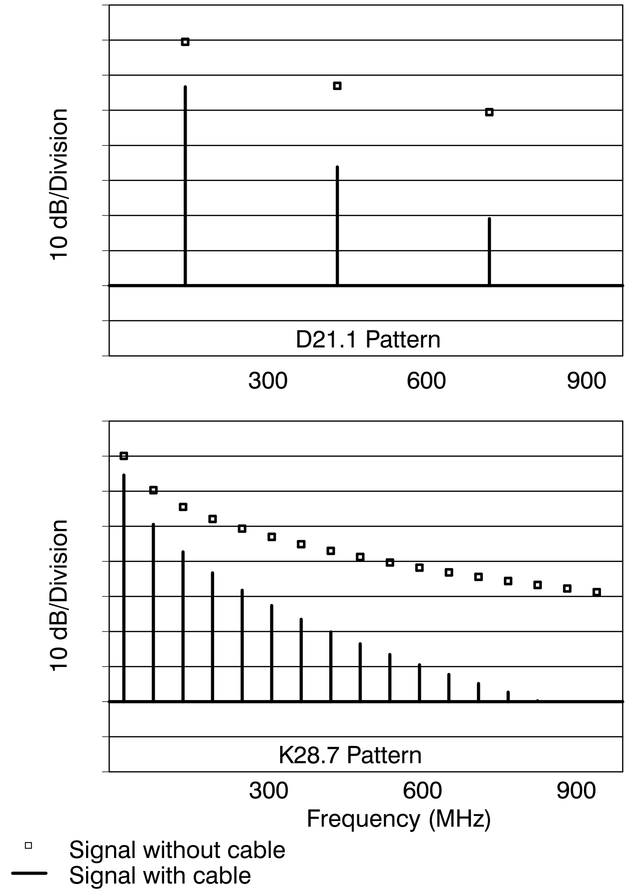

Figure 5. Spectrum of Synthesized D21.5 and K28.7 Patterns After 100m of RG59 Cable

With these data rate and cable combinations, only 25% of the peak-to-peak amplitude of the D21.5 (1010101010) pattern remains after 100 meters of cable, while the K28.7 (1111100000) pattern has nearly 60% of its signal available.

Figure 7 shows the actual measured signals at the source and after 100 meters of cable. While the measured amplitudes are a close match to the calculated amplitudes, the waveshape of the K28.7 signal at the end of the cable is significantly different. The cause of this distortion is a variation in propagation velocity versus frequency known as dispersion.

Dispersion

Dispersion is a propagation characteristic more commonly linked to optical fibers. This causes light at different wavelengths to propagate at different rates through the fiber.

This same phenomenon exists in copper cables where higher frequency signals propagate faster than slower frequency signals. This variation in propagation is caused by two different phenomena: a change in dielectric constant of the cable dielectric with frequency, and a change in the reactance of the cable with frequency.

Dielectric Dispersion

Recall from the "Driving Copper Cables with HOTLink" application note that for coaxial cables and stripline transmission lines

$$V_P = \frac{1}{\sqrt{\varepsilon_r}} \qquad \text{Eq. 12}$$

If the dielectric constant (ε_r) for a transmission line remains constant across all frequencies, the signal

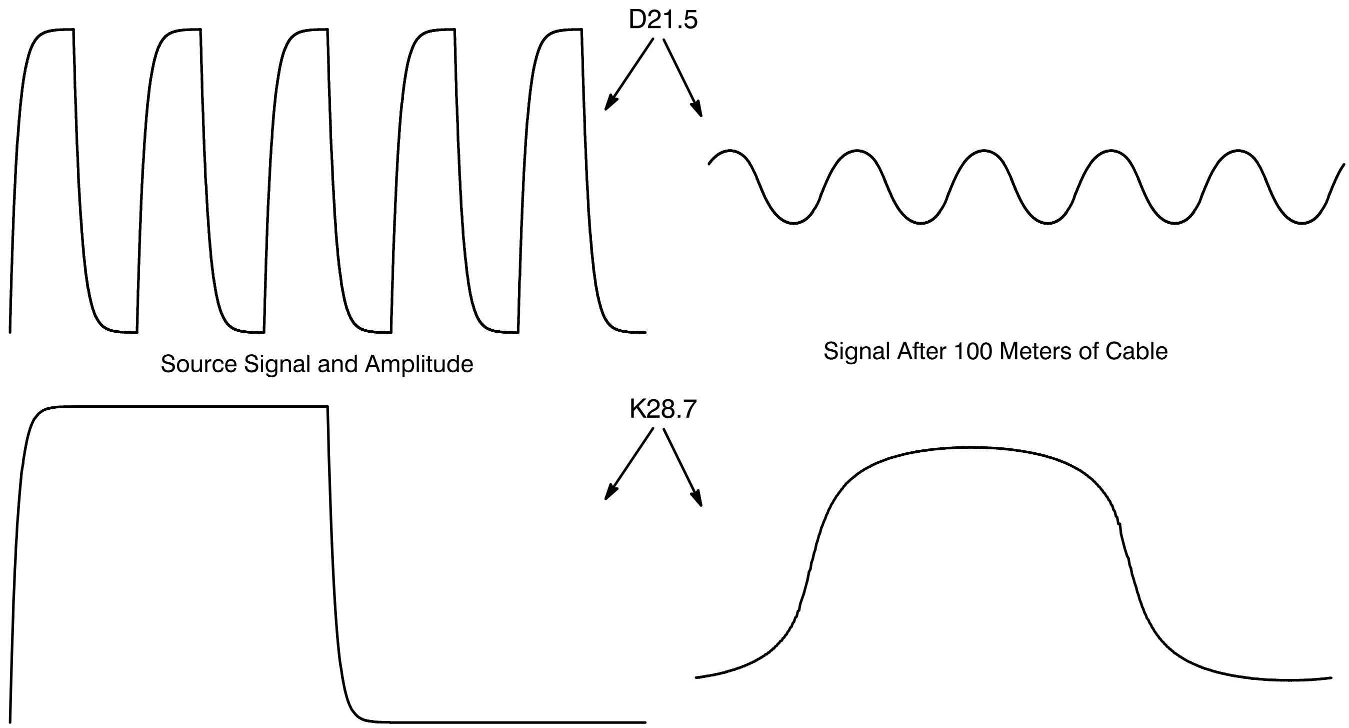

Figure 6. Synthesized D21.5 and K28.7 Waveforms with Simulated Cable Attenuation

spectral components will propagate down the transmission line at the same rate. Unfortunately, many dielectrics are not stable with frequency. Dielectrics such as bakelite, glass, rubber, and PVC (polyvinyl chloride) exhibit from several percent to 10s of percent change in dielectric constant over the 1-MHz to 1-GHz frequency range. Common circuit board materials also are not stable with frequency. *Figure 8*

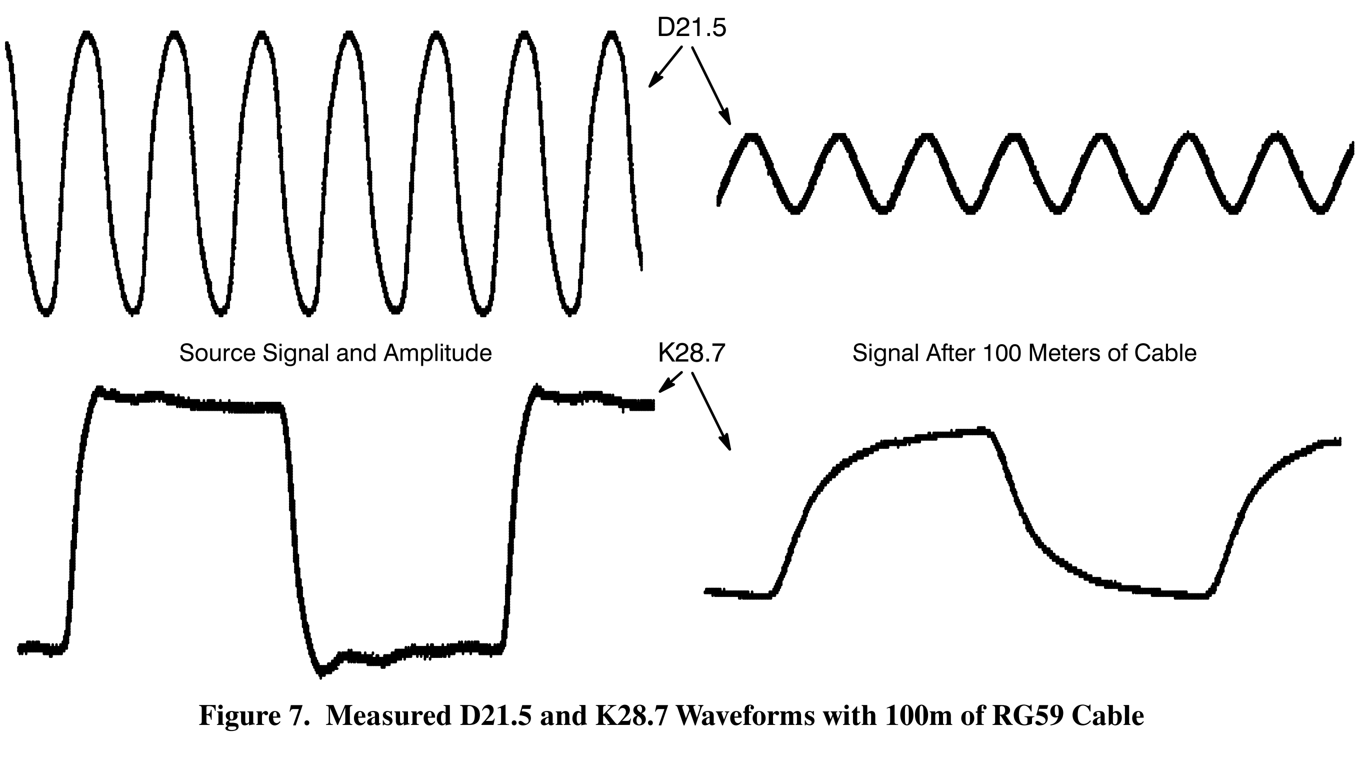

Figure 7. Measured D21.5 and K28.7 Waveforms with 100m of RG59 Cable

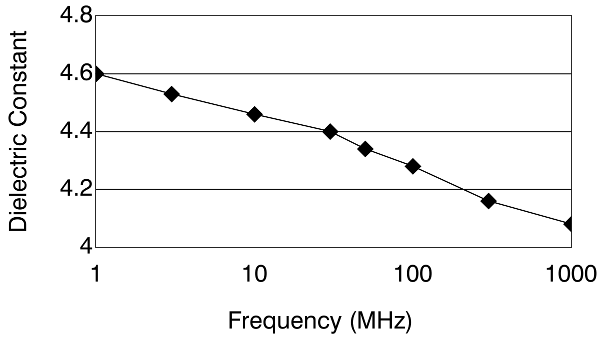

Figure 8. Dielectric Constant of G10/FR4 Circuit Board Laminate

shows how the dielectric constant (ε_r) changes in G10/FR4 circuit board laminate (Reference 8).

Applying these ε_r values to the spectral components in the K28.7 and D21.5 signals shown in *Figures 5* and *6* yields signal components traveling at the rates listed in *Table 3*. When sending an actual data stream, these and other signal components are present in the transmission line at the same time. The D21.5 fundamental (150 MHz, equal to the bit-rate) has a wavelength in the transmission line of only 0.974 meters. Because the K28.7 fundamental (15 MHz, equal to the byte-rate) is traveling 1.3% slower than the D21.5 fundamental, this signal component will lag the D21.5 fundamental by 90° of phase (equal to 50% of one bit time) after only 16.2 meters of transmission line.

Table 3. D21.5 Signal Propagation Rates

Frequency	ε_r	Vp (%)	Vp (m/second)
15 MHz	4.45	47.4%	1.422×10^8
45 MHz	4.35	47.9%	1.438×10^8
150 MHz	4.22	48.7%	1.460×10^8
450 MHz	4.12	49.3%	1.478×10^8

The 90° phase point was selected because it is equivalent to 100% jitter in the received bits. In reality other signals components will cross this 90° phase point in much less than the 16 meter length. This is because a normal data stream contains other signal components both higher and lower in frequency than the two selected here. Since these other signals components are traveling both slower and faster than the 15 MHz and 150 MHz signals used in this example, the 90° point will be reached with a much shorter transmission line. Due to the limited energy present in each of these signal components, they individually cannot close up the received signal eye.

Other Dispersion Factors

Good RF-grade cables are usually made with stable dielectrics; i.e., those that exhibit only minor changes in dielectric constant over frequency. These cables are usually constructed using dielectrics based on polyethylene, polypropylene, polyolefin, polystyrene, and various Teflon® derivatives. These dielectrics vary in dielectric costant by less than 0.5% from 100 Hz through 10 GHz. Calculations for dielectric based dispersion show little interaction even after hundreds of meters of cable, yet these cables still exhibit dispersion. The dispersion effect in these cables is caused by the variation in reactance that occurs in the cable with changing frequency. The dispersion caused by reactance (per unit length of cable) is much smaller than that caused by non-frequency stable dielectrics. This allows cables based on stable dielectrics to be used for much longer signal transmission.

In reality the calculation of velocity of propagation in *Equation 12* is a simplified form that only assumes first order effects. A proper calculation must take into account all four distributed properties (R, G, C, and L) in a transmission line. This is normally described as the complex propagation constant γ, and is calculated using *Equation 13* (Reference 4).

$$\gamma(\omega) = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \qquad \text{Eq. 13}$$

The complex propagation constant γ consists of a real portion α, representing the attenuation of the signal, and an imaginary portion $j\beta$, representing the angular velocity of the signal. Both of these are per unit length of the transmission line. This shows that the propagation rate is based on all four parameters, not just the dielectric constant of the line.

With RF-grade dielectrics the dielectric constant remains stable over frequency and thus does not effect the propagation rate. The conductance parameter does increase at a rate directly proportional to frequency. At frequencies over a few megahertz its effect, relative to that of $j\omega C$, is so small that it is usually discarded. With the long transmission lines considered here this very small effect is still important. The resistance parameter also increases with frequency. This change in resistance is caused by the previously described skin effect, where the transmission line resistance is affected by the uneven current distribution.

The distributed inductance is also affected by frequency. The total inductance is a sum of the external inductance (that present between the two conductors of the transmission line) and the internal or self inductance of the conductors (assuming a dielectric free of magnetic properties).

Each of these pieces has a small effect on the total propagation rate of signals. Most of these effects are only observable with cables extending for tens or hundreds of meters.

Equalization

Equalization is the application of frequency selective gain or attenuation to compensate for distortion. Equalization is used in analog audio, analog video, and digital signal transmission systems to compensate for characteristics of the system and the distortion created by the operating environment.

For HOTLink-based communications, the primary cause of signal distortion is the non-linear characteristics of the interconnecting cable. This cable attenuates the high-frequency signal components much more than the low-frequency signal components, and introduces a frequency selective phase delay into the signal.

As the length of the interconnecting cable increases, so does the signal distortion. At some point the distortion becomes so great that the HOTLink receiver is no longer able to correctly recover the serial datastream. While there is still sufficient amplitude available in the signal, the data-dependent jitter (DDJ) exceeds the jitter tolerance of the HOTLink receiver (typically >90%). To allow reliable communications with these long cables it is necessary to “equalize” the the cable.

Equalization Circuits

Equalization can take many forms. For many low-frequency circuits, equalization often uses a combination of active and passive components to create frequency selective filters that provide specific amounts of gain or attenuation for a signal. These same filters may be made to automatically adapt to different cable, frequency, and distance combinations.

At higher operating frequencies (such as those used with HOTLink), the design and implementation of active filters becomes more difficult, and equalization is usually performed using only fixed passive components, followed by a non-frequency-selective amplifier. This provides the lowest cost form of equalization, but is not as flexible as an adaptive/active equalization circuit.

With a passive equalizer, the only functions that the circuit can provide are attenuation and phase change—they cannot provide gain (peak amplitude of some signals may increase, but this is due to alignment of the signal component phasors). To equalize a copper cable, the circuit must operate in a manner opposite that of the interconnecting cable. This effectively means a high-pass filter that delays the phase of high-frequency signal components.

Many such circuits are available, all with different topologies and characteristics. A simple equalizer circuit recommended for HOTLink use is explained in detail in the following example.

Equalizer Example

A pair of equalizers suitable for use with HOTLink are shown in *Figure 9*. The Bridged-H circuit is a balanced circuit that operates with balanced transmission lines. This balanced equalizer may also be used with unbalanced cables if placed on the balanced side of a balun coupling transformer.

The Bridged-T circuit is an unbalanced form of the Bridged-H equalizer. This circuit is designed for use with unbalanced transmission lines. When used with a HOTLink receiver that is transformer coupled, this circuit must be used in the unbalanced portion of the transmission line. It may be used with coaxial (or other unbalanced) cables by placing the

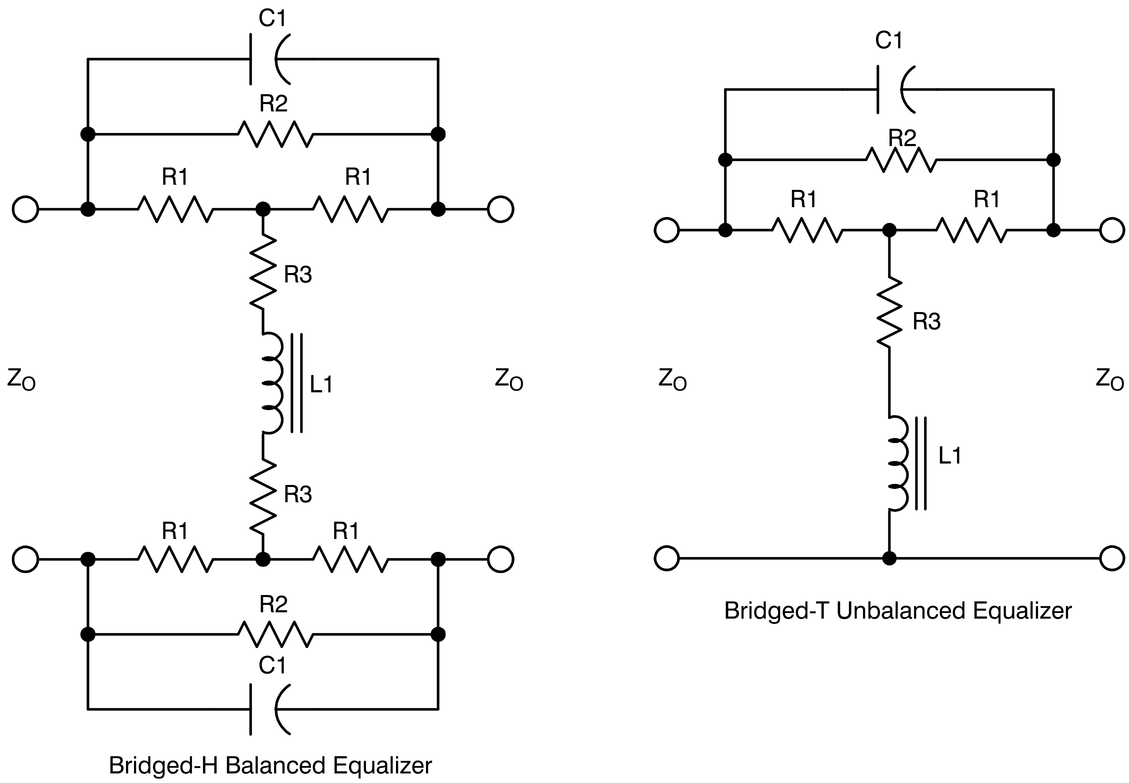

Figure 9. Constant Impedance Equalizer Circuits

circuit between either end of the transmission line and the coupling transformer.

Both of these circuits are AC-forms of a fixed-attenuator or "pad". A pad is often used for impedance matching or attenuating between a source and destination, with minimal parts count and minimum loss. The equalizers in *Figure 9* are converted to their pad equivalent by removing the capacitors and shorting out the inductor. Unlike some pads which can perform impedance transformation, these Bridged-H and Bridged-T circuits require the input and output impedances to be the same.

These equalizers, when properly implemented, appear across a wide frequency range as a DC resistance at the end of a cable. For frequencies at or near DC, the gain (insertion loss) is determined only by the resistors. As the frequencies approach the active region of the filter, the reactive nature of the capacitor starts to have an effect. The higher frequencies see less reactance and are passed through the capacitor with minimal attenuation. The inductor is selected to exactly match (but with increasing reactance) the frequency response characteristics of the capacitor(s).

The component values for these circuits are determined by the specific cable type selected, the frequency of operation, and the desired distance of operation. The design equations for both structures are detailed in *Table 4*. Because the balanced Bridged-H circuit is based on the unbalanced Bridged-T (and all values for it may be derived from the Bridged-T equations), only the Bridged-T circuit will be explained in detail.

Table 4. Equalizer Equations

Component	Bridged-T	Bridged-H
R1	Z_O	$Z_O/2$
R2	Z_O*X	$(Z_O*X)/2$
R3	Z_O/X	$Z_O/(2*X)$
C1	$L1/(Z_O^2)$	$(2*L1)/(Z_O^2)$
L1	$C1*Z_O^2$	$(C1*Z_O^2)/2$

Z_O = characteristic impedance of cable,
X = see *Equation 15.*

Equalizer Example

The R1 value is the easiest to determine. For the Bridged-T circuit it is equal to Z_O. For the RG59 cable documented previously (Z_O=75Ω), the R1 value would be 75Ω.

The relationship for R2 and R3 determines both the DC-gain (loss) of the equalizer and the correction attenuation slope. To keep a constant impedance, it is necessary for

$$Z_O = \sqrt{R2 \cdot R3} \qquad \text{Eq. 14}$$

The gain is determined by the ratio of each resistor to the filter impedance, and a gain constant X. The gain constant (X) determines how much insertion loss the filter should have at low (near DC) frequencies, and is determined using *Equation 15*.

$$X = \sqrt{10^{\left(\frac{dBAttenuation}{10}\right)}} - 1 \qquad \text{Eq. 15}$$

Attenuation Slope

This same gain constant also determines the slope of the attenuation curve in the active region of the filter. For equalization purposes the gain constant must be determined by the slope of the transmission line attenuation over the main frequency range of interest.

The transmission line presents an attenuation verses frequency slope that increases with cable length. *Figure 2* shows that the source (cable) attenuation function is linear when plotted in log/log space (attenuation verses frequency). To flatten the system frequency response the equalizer must then present an attenuation verses frequency slope that is equal in magnitude but opposite in slope to that of the cable.

Unfortunately a single pole filter (like that used here) can only generate a correction slope of at most −20 dB/decade. The source signal attenuation also increases at a logarithmic rate per decade rather than a linear rate per decade. This means that the correction applied to the signal can only be a coarse approximation rather than a perfect correction.

Using the RG59 cable documented earlier, and assuming a cable length of 100 meters and a data rate of 300 Mbaud, it is possible to calculate the approximate attenuation slope (in dB/decade) that the equalizer must attempt to correct. The goal is to have the low-frequency content of the received signal match the high-frequency content at a specific length of cable.

The data from *Table 2* identifies that the attenuation at 150 MHz (the bit-rate equivalent sinusoidal frequency of 300 Mbaud) is 12.8 dB for a 100 meter cable. At the 30 MHz frequency (the byte-rate equivalent sinusoidal frequency) the attenuation is 5.4 dB. These two points are then used to determine the necessary correction attenuation slope (in dB/decade) using *Equation 16*. Entering these values into *Equation 16* yields an attenuation slope of 10.61 dB/decade.

$$slope = \frac{A1 - A2}{log(F1) - log(F2)} \qquad \text{Eq. 16}$$

Equalization Slope

To equalize the cable it is necessary to present a correction having a matched slope but starting from the bit-rate fundamental frequency. This slope is controlled only by the R2/R3 resistors, with the frequency being determined by C1/L1. As the R2/R3 resistor ratio varies (as set by the gain constant X) the attenuation slope varies from between zero and 20 dB/decade. The necessary gain constant may be determined directly using *Equation 17*. Using the previously calculated source slope yields a gain constant of 2.224.

$$X = \sqrt{3.9 \times \left[\tan\left(slope \times \frac{\pi}{40}\right)\right]^{2.49}} \qquad \text{Eq. 17}$$

Note: This equation was derived from empirical data. Its function matches simulated response curves to within 0.15 dB for the entire 0 to 20 dB/decade range.

With the gain constant now available, the values of R2 and R3 may be determined. Using the equations from *Table 4* for R2 and R3, these calculate to R2=166.8Ω and R3=33.7Ω. Inserting this same gain constant into *Equation 18* sets a DC gain of −10.17 dB.

$$dBattenuation = 10 \times \log[(X + 1)^2] \qquad \text{Eq. 18}$$

Center Frequency

The L1 and C1 components are used both to select where the signal attenuation occurs, and to keep the equalizer impedance constant. To maintain the a constant impedance in the equalizer, the product of the shunt and bridge impedances must always equal the square of the characteristic impedance. In terms of L1 and C1 this can be reduced to the relationship in *Equation 19*.

$$Z_O = \sqrt{\frac{L1}{C1}} \qquad \text{Eq. 19}$$

Setting the roll-off point for the high-pass filter is not quite as intuitive. At first glance the equalizer appears as a single-pole filter yielding a fixed 6 dB/octave or 20 dB/decade attenuation below a cutoff frequency. This is the actual filter response when set for a DC gain of 0 (DC loss= ∞) by removing R2 and shorting R3. In this configuration the −3 dB cutoff frequency is determined using *Equation 20*.

$$f_C = \frac{1}{2\pi\sqrt{L1 \cdot C1}} \qquad \text{Eq. 20}$$

Adding R2 and R3 back into the circuit however changes the slope of the attenuation curve, moves the upper cutoff frequency, and adds a lower cutoff frequency point. *Figure 10* shows the gain and phase response for this equalizer implemented with an arbitrarily selected (but properly balanced) C1/L1 pair of 200 pF and 1125 nH. The attenuation slope is correct, but the location within the frequency spectrum is not. An examination of the phase response curve shows that it peaks at the midpoint of the active region of the filter.

The capacitor C1 is responsible for the location of the attenuation curve within the frequency spectrum. As the capacitance is decreased, the curve is shifted higher in frequency, but with an identical slope. The correct capacitor (and corresponding inductor) are selected when the line determined by the equalizer attenuation slope intersects the bit rate frequency (150 MHz for this example) at 0 dB.

Unfortunately, any simulation or measurement will show that the attenuation slope is not linear at the upper and lower ends of the active region of the filter. The only point on the gain curve whose slope actually matches the desired correction slope is at the midpoint of the curve, located at the same frequency

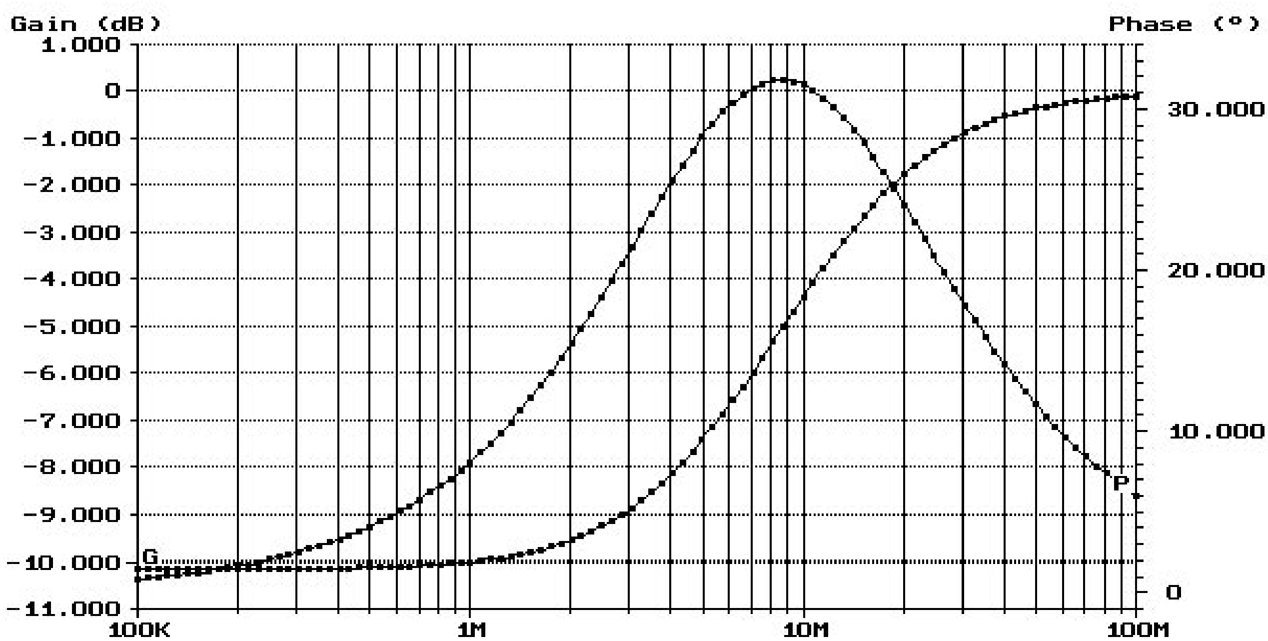

Figure 10. Gain/Phase Plot for Initial C1/L1 Values

as the peak in the phase response (8.5 MHz). The attenuation at this point is exactly half the DC attenuation (−5.08 dB).

The filter response of the present circuit is obviously too low for proper compensation of a 300 Mbaud data stream. What is necessary is to shift this midpoint to a different frequency. This new midpoint intercept frequency is calculated using *Equation 21*. Using this equation with the current bit-rate frequency (150 MHz), DC gain (−10.17 dB), and equalization slope (−10.61 dB/decade) yields a new center frequency of 49.8 MHz.

$$F\_new = 10^{\left(\log(F\_bit\_rate) - \frac{DC\_Gain/2}{slope}\right)} \qquad \text{Eq. 21}$$

To determine the correct C1 and L1 values that will center the filter response through this point requires determining the magnitude of the reactance phasor at this point. The reactance at this center point in the filter response remains the same with any properly matched C1/L1 pair. In the gain/phase plot in *Figure 10*, the center frequency is at 8.5 MHz. The impedance phasor magnitude for the bridge (R2/C1) and shunt (R3/L1) paths are calculated using *Equations 22* and *23* respectively.

$$X_B = \frac{1}{\sqrt{\frac{1}{R2^2} + (2\pi f \cdot C1)^2}} = 81.6\Omega \qquad \text{Eq. 22}$$

$$X_S = \sqrt{R3^2 + (2\pi f \cdot L1)^2} = 68.9\Omega \qquad \text{Eq. 23}$$

These X_B and X_S values are the magnitudes of the complex impedances present in the R2/C1 and R3/L2 component pairs respectively. Solving for the specific C1 and L1 components at the desired 49.8 MHz midpoint frequency involves converting the impedance vectors into their real and imaginary components, and determining what size component will yield the proper reactance at the specified center frequency. The calculations for C1 and L1 are shown here in *Equations 24* and *25*.

$$L1 = \frac{\sqrt{X_s^{\ 2} - R3^2}}{2\pi f} = 192.2\ mH \qquad \text{Eq. 24}$$

$$C1 = \frac{\sqrt{\frac{1}{X_B^{\ 2}} - \frac{1}{R2^2}}}{2\pi f} = 34.2\ pF \qquad \text{Eq. 25}$$

Placing these new C1 and L1 components into the Bridged-T equalizer yields the filter response shown in *Figure 11*. The slope of the curve (in dB/decade)

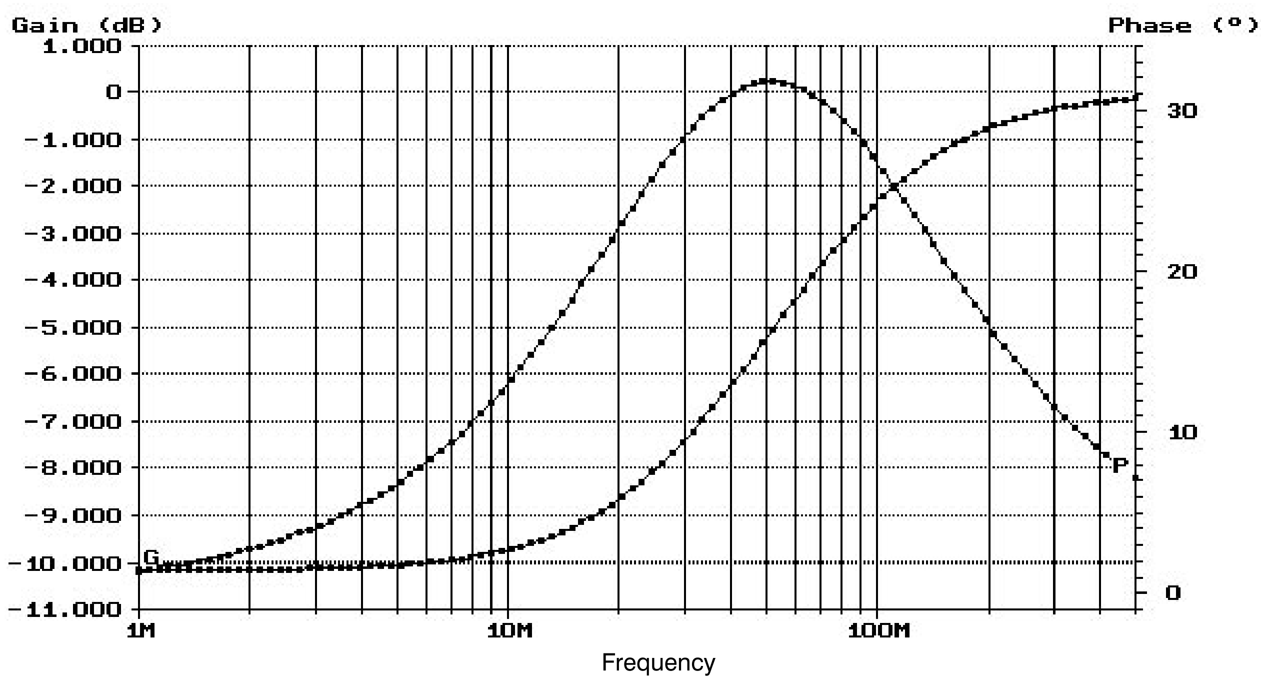

Figure 11. Gain Phase Plot for Final C1/L1 Values

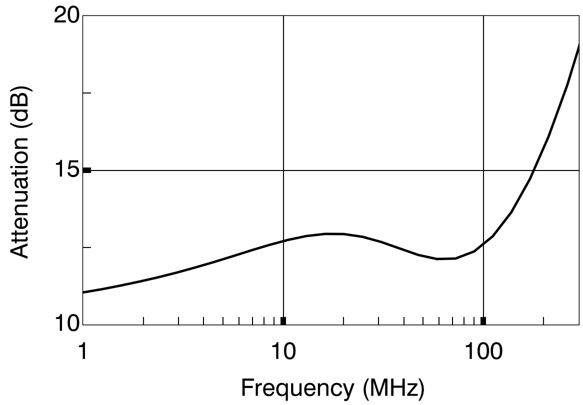

Figure 12. Combined Cable and Equalizer Attenuation

remains the same, but now the phase response peak occurs near 50 MHz.

Composite Response

Figure 12 shows how close this equalization matches the cable's frequency response. This curve is a sum of the cable and equalizer attenuations at each frequency point. Note that the link response (100 meters of cable and the equalizer) does not vary by more than 2 dB for over two decades of frequency spectrum. Once the signal spectral components are above the bit-rate frequency of the filter, the cable attenuation becomes dominant and the attenuation slope increases dramatically. Slight alterations of the equalizer slope and frequency intercept can modify this curve to meet specific frequency response and flatness requirements.

Implementation Constraints

While the numeric calculations allow a design to be implemented on paper, bring such a design into the real world is much different. Finding components with even 1% accuracy can be difficult if not impossible. Parasitic reactances present in any component also effect the response of the equalizer circuit. This means that even the best equalizer will wind up being a number of compromises.

Resistors

The selection of resistor values is probably the easiest to make. These components are available in wide ranges of values and tolerances. For most equalizer implementations, these parts should be 1% tolerance components.

Because of the wide frequency range that the equalizer must cover, care should also be exercised in the selection of the type of resistive element used. Carbon composition and carbon film resistors have significant capacitive parasitics and should not be used in sizes over 100Ω in equalizers of this type. A better choice here would be metal film resistors.

The physical size of the component also makes a difference. Generally the smaller the components physical size, the lower the inductive and capacitive parasitics present.

Inductors

The inductor is the most difficult component to select, primarily because they are manufactured in so few standard sizes. In the range from 10 nH through 2000 nH (the range most likely to be used with HOTLink) all manufacturers provide the same series of part values in each decade of size. These values are 10, 12, 15, 18, 22, 27, 33, 39, 47, 56, 68, and 82. All other standard sizes are found by multiplying these values by 10, 100, 1000, etc. Custom sizes are available from some manufactures, but generally at a significant cost difference.

Another problem that plagues most inductors is a low series resonant frequency. For the equalizer to operate correctly (within its designed range of operation), the inductor must continue to provide increasing amounts of reactance with increasing frequency. This means making sure that the series resonant frequency of the inductor is greater than the bit-rate frequency of the data stream. The best inductors for this are generally made from a multi-layer ceramic construction.

The last concern is manufacturing tolerance. Unlike resistors where 1% tolerance parts are low in cost and widely available, the common tolerance for inductors is 10%. A few manufacturers also offer 5% and 2% tolerance parts.

Capacitors

The choice of capacitors is almost dictated by the available sizes of inductors, and the small quantity of capacitance required for most equalizers. This will generally fall in the 10 to 200 pF range. The majority of all chip capacitors in this range are made with a temperature stable low-K dielectric known as either NP0 or C0G. Other high-K dielectrics should not be used, both for their instability over temperature and for the ferroelectric effect these high-K dielectrics exhibit.

While capacitors also have a series resonant frequency, it is not generally a concern when using the types and sizes of capacitors required for these equalizers. In almost all cases the series resonant frequency is well above the bit-rate frequency and therefore of only minor concern.

Board Layout

Just as incorrect component selection can greatly effect the frequency response of an equalizer, so can a poorly implemented layout. The circuit traces, pads, and vias all have an effect on the circuit operation. The following guidelines should be applied to minimize these effects.

- Use as short of traces as possible to minimize the trace inductance and capacitance.
- Keep all components in close proximity to each other.
- Minimize the number of vias. These structures can be routed on a single layer without vias.
- For the Bridged-H balanced equalizer, keep routing symmetrical to keep the parasitics balanced.

Conclusion

Communications on electrically long transmission lines are possible with many types of media. How far a signal may be reliably transmitted is a function of many driver, cable, filter, and receiver characteristics. Application of equalization filters can allow communication over distances well beyond that of non-equalized systems. These equalizers may be implemented with a minimal number of low cost passive components.

References

1. True, Kenneth M., *Long Transmission Lines and Data Signal Quality*, AN808, National Semiconductor
2. Orr, William I., *Radio Handbook*, 23rd Edition, SAMS, 1992
3. Belden Master Catalog, Cooper Industries, Inc., 1992
4. True, Keneth M., *Data Transmission Lines and Their Characteristics*, AN−806, National Semiconductor
5. True, Keneth M., *Long Transmission Lines and Data Signal Quality*, AN−806, National Semiconductor
6. *Fibre Channel Standard*, ANS X3.230-1994, American National Standards Institute, 1994
7. Ramierez, Robert W., *The FFT, Fundamentals and Concepts*, Tektronix, Inc. 1985
8. AdCore Product Anouncement, GIL Copper Clad Laminates, Alpha Corporation, 1995

HOTLink is a trademark of Cypress Semiconductor.
Teflon is a registered trademark of DuPont.