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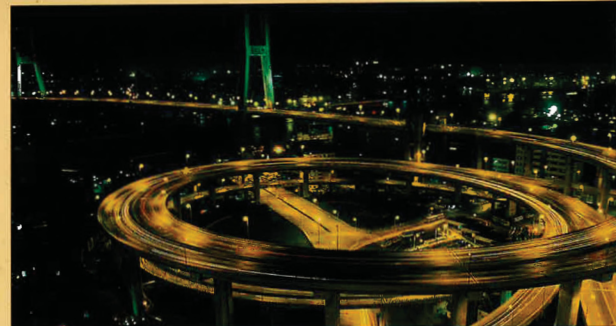
DSL

*Simulation Techniques and Standards
Development for Digital Subscriber
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MACMILLAN TECHNOLOGY SERIES

DSL

*Simulation Techniques and Standards
Development for Digital Subscriber
Line Systems*

Dr. Walter Y. Chen

DSL: Simulation Techniques and Standards Development for Digital Subscriber Line Systems

By Dr. Walter Y. Chen

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About the Author

Dr. Walter Y. Chen was born in Shanghai, China, on September 8, 1956. He received a B.S. (summa cum laude) degree in electrical engineering from the Polytechnic Institute of New York, Brooklyn, in 1982, an M.S. degree in electrical engineering from California Institute of Technology, Pasadena, in 1983, and a Ph.D. in electrical engineering from Polytechnic University of New York in 1989.

From 1982 to 1987, he was a Member of Technical Staff at AT&T Bell Laboratories, where he worked on PBX systems engineering and voice band modem VLSI chip design. From 1987 to 1989, he was a Member of Technical Staff at NYNEX Science and Technology, where he conducted computer network traffic and adaptive signal processing simulation studies. From 1989 to 1995, he was a research scientist at Bellcore, where he was involved with HDSL, ADSL, and HFC modem simulation studies and standards development. From 1995 to 1997, he was a research branch manager at the DSPS R&D Center of Texas Instruments, Inc., for the development of DSL technologies and High-throughput Digital Home Networks.

Dr. Chen is currently with the Information Systems Group of Motorola working on next-generation DSL and Digital Home Network technologies. Since 1989, he has made more than 40 technical contributions at T1E1.4, ADSL Forum, and IEEE 802.14 standards committees. His main research interest is the application of circuit and system theories to the area of advanced digital communication. He holds four U.S. patents and has several others pending.

Dedication

To Aaron and Brian, and to Nancy and our parents.

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First, I would like to thank all the folks at Macmillan Technical Publishing for making this book a reality. Especially, I would like to thank Jim LeValley for getting me started with his inquiring email message. I also would like to thank Amy Lewis and Tracy Hughes for their assistance, guidance, and patience during the course of this project.

The transfer function has the same shape as that of a single-ended hybrid; the loss difference is due to the scaling factors of $1/2$ over $1/3$ when $R_1^d = 2R_1^s$. The loss difference is $20 \times \log_{10}(3/2) = 3.52$ dB. The performance of a differential hybrid, however, is the same as that of a single-ended hybrid circuit because the received signal through a differential hybrid circuit also experiences 3.52 dB more loss.

End Notes

1. Midcom web page at <http://www.midcom-inc.com/technotes/tn17.html>.
2. W. Y. Chen, J. L. Dixon, and D. L. Waring, "HDSL Echo Cancellation," *IEEE JSAC*, vol. 9, no. 6 (August 1991): 848–860. G. S. Moschytz and S. V. Ahamed, "Transhybrid Loss with RC Balance Circuits for Primary-Rate ISDN Transmission Systems," *IEEE JSAC*, vol. 9, no. 6 (August 1991): 951–959.

CHAPTER

7

Analog Front-End Precision

Some transmission systems can be realized with analog circuit technology only, such as transmit line driver, receiver amplifier, and threshold detector. The transmission potential cannot be fully realized, however, with the use of current analog circuit technology only. Digital processing techniques, such as adaptive channel equalization and adaptive echo cancellation, have played a very important role at maximizing the transmission efficiency of DSL systems.

On the other hand, the transmit signal still needs to be in analog format to fit the analog twisted-pair telephone loop better. Information after digital signal processing with limited precision is converted to analog format through a *Digital to Analog Converter* (DAC), and the received analog signal is converted through an *Analog to Digital Converter* (ADC) for further digital signal processing.

The precision of DAC and ADC should be fine enough to realize the potential of the analog telephone subscriber loop channel. In other words, the precision should be chosen such that the DAC and ADC quantization noise will not be a limiting factor for the transmission performance of DSL systems. The required DAC and ADC precision depends on the following:

- The transmit signal line code
- The transmit shaping filter
- The characteristics of the channel
- The attenuation of the echo path

There are baseband, passband single-carrier, and multiple-carrier line codes used for DSL systems. The four-level baseband line code, also called 2B1Q for 2 bits per quaternary, has been used for both the *Basic Rate Access ISDN channel* (also called DSL) and for HDSL.

A baseband line code usually consists of many parity symmetric and equally spaced signal levels. Passband single carrier line codes, such as *Quadrature Amplitude Modulation* (QAM) and *Carrierless AM/PM* (CAP), have also been proposed for DSL systems. A passband single carrier line code has a two-dimensional signal. The two-dimensional signal is divided into an in-phase channel and a quadrature channel implemented by the cosine and sine phases of the single carrier, respectively.

Parity symmetric and equally spaced multiple signal levels are used for both in-phase and quadrature channels. The *Discrete MultiTone* (DMT) line code is the standard for ADSL. A DMT line code consists of a multiple of single carriers implemented through the *Discrete Fourier Transform* (DFT). The number of carriers available to carry information is equal to, or slightly less than, half of the DFT size. Each carrier of a DMT line code has the same two-dimensional signal as that of a passband single carrier system.

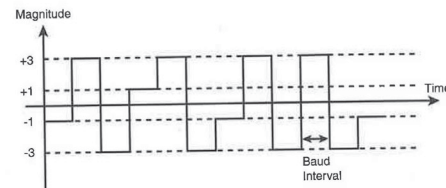
Different line codes have different peak-to-average voltage ratios. The peak-to-average voltage ratio can also be affected by transmit shaping filters for a DAC and by the channel and the echo path for an ADC. This chapter discusses issues related to the determination of DAC and ADC precision requirements including the following:

- **Peak-to-average ratio (PAR) of line codes.** PARs are calculated for baseband multiple level, passband single carrier, and DMT multiple carrier line codes. Relationships of PARs to the number of signal levels, constellation size, and the number of carriers are derived. The effect of clipping for the DMT line code is also discussed.
- **Effects of transmit filter, channel, and echo path.** The use of a transmit filter generally leads to a higher PAR. The PAR can also be increased for a signal passing through the telephone subscriber loop channel. The echo path could produce the same effect.
- **Characteristics of ADC and DAC.** The effect of ADC and DAC quantization noise and its relationship to required PAR are discussed. The effect of ADC and DAC nonlinearity is compared with that of quantization noise.
- **ADC and DAC precision requirements.** Combining PARs of different line codes and the quantization noise of ADC and DAC, precision requirements are estimated for a few practical DSL system examples.

7.1 Transmit Signal Peak-to-Average Voltage Ratio

A binary baseband signal can have signal levels of -1 and $+1$. A four-level baseband signal can have signal levels of -3 , -1 , $+1$, and $+3$. Figure 7.1 shows the waveform of a baseband 2B1Q line code.

Figure 7.1 A Baseband 2B1Q Line Code



For the previous and following baseband signals, the distance between adjacent signal levels is maintained to be 2. In general, an n level baseband signal can have signal levels of $-n+1$, $-n+3$, ..., $n-3$, $n-1$, for which n is an even number.

Assuming a load resistance of one, the average power for a binary signal is one. Using the same load resistance value assumption, the average power for a four-level signal is $(1^2 + 3^2)/2$. In general, the average power for an n level baseband signal is

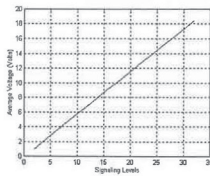
$$\frac{2}{n} (1^2 + 3^2 + \dots + (n-1)^2 + (n-1)^2) = \frac{n^2 - 1}{3}$$

Assuming a load resistance of one, the average voltage of this n level baseband signal is

$$\sqrt{\frac{n^2 - 1}{3}}$$

Figure 7.2 shows average voltage levels for different levels of baseband signals assuming a voltage of two volts between adjacent signal levels.

Figure 7.2 Average Signal Magnitude of Baseband Signal



For a peak voltage level of $(n-1)$ volts, the peak-to-average voltage ratio of an n level baseband signal is

$$(n-1) \sqrt{\frac{3}{(n^2-1)}} = \sqrt{\frac{3(n-1)}{n+1}}$$

Figure 7.3 shows peak-to-average voltage ratio (PAR) for baseband signals with different numbers of signaling levels.

Figure 7.3 Peak-to-Average Voltage Ratio of Baseband Signal

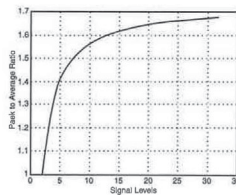
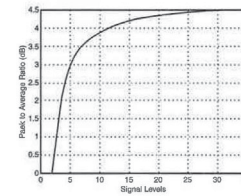


Figure 7.4 shows the same PAR in terms of dB.

Figure 7.4 Peak-to-Average Voltage Ratio (dB)



In practice, the peak transmit voltage is limited by the voltage level of a power supply. For a 5 volt power supply, the peak-to-peak voltage can be as high as 4.6 volts without introducing too much amplifier nonlinearity. Using a single-ended hybrid circuit, the peak signal level can be represented by a voltage level of 2.3 volts. Under the power supply voltage limited condition, the average transmit power for an n level baseband signal with a load resistance of 135 ohms is then

$$P_T = \frac{2.3^2}{135} \left(\frac{n+1}{3(n-1)} \right)$$

For an example of a four-level signal, the average transmit power is

$$P_T = \frac{2.3^2}{135} \left(\frac{5}{3 \times 3} \right) = 0.02177 \text{ Watts} = 13.38 \text{ dBm}$$

Because of the increased PAR, the average transmit power for a 32-level baseband signal with the same 5 volt power supply is 2 dB lower.

7.1.1 PAR of Passband Single Carrier Line Codes

A two-dimensional passband single carrier line code can be characterized by its constellation. A CAP line code is a little different compared with the QAM line code. The difference is explained in Chapter 8, "Channel Equalization." CAP and QAM line codes can, however,

Figure 8.41 Channel Equalizers for a 1 Mile, 26-Gauge Loop

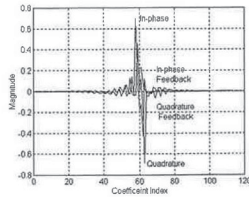


Figure 8.42 Channel Equalizers for a 2 Mile, 26-Gauge Loop

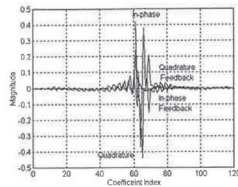
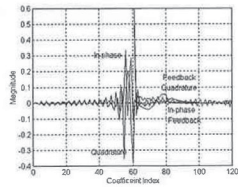


Figure 8.43 Channel Equalizers for a 3 Mile, 26-Gauge Loop



Because FBFs does not contribute to the noise enhancement, MMSEs for a CAP DFE system are the same as those described for the CAP linear equalizer system. The additional performance improvement comes from a different set of CAP FFF filter coefficients:

$$\xi_{\text{in},\min} = E[a_k^2] - \begin{bmatrix} R_{Xa}^{1T} & R_{Xa}^{2T} & R_{Xa}^{3T} \end{bmatrix} \begin{bmatrix} Q_{i,\text{opt}}^1 \\ Q_{i,\text{opt}}^2 \\ Q_{i,\text{opt}}^3 \end{bmatrix}$$

$$\xi_{\text{qu},\min} = E[b_k^2] - \begin{bmatrix} R_{Xb}^{1T} & R_{Xb}^{2T} & R_{Xb}^{3T} \end{bmatrix} \begin{bmatrix} Q_{q,\text{opt}}^1 \\ Q_{q,\text{opt}}^2 \\ Q_{q,\text{opt}}^3 \end{bmatrix}$$

Equalizer output SNRs for QAM and CAP systems are summarized in Table 8.2 for different equalizer structures under different receiver front-end noise power density levels for the 2 mile, 26-gauge twisted-pair loop. Results show that DFE performance is only slightly better for the passband system. However, a QAM equalizer with a fractionally spaced FFF could make performance robustness irrelevant to the sample timing phase.

Performances of a QAM system and a CAP system with fractionally spaced FFFs should be about the same. The difference in the table is caused by the sample timing phase of the channel because the QAM system used a two times over-sampled FFF and the CAP system used a three times over-sampled FFF.

Table 8.2 Passband Equalizer Output SNRs

Noise Level	QAM Linear	QAM DFE	Fractional QAM Linear	CAP Linear	CAP DFE
-110 dBm/Hz	22.0	22.2	25	26.0	26.8
-120 dBm/Hz	32.0	32.2	35	36.0	36.8
-130 dBm/Hz	42.0	42.2	45	45.9	46.7

8.4 Equalizers for Discrete Multitone Systems

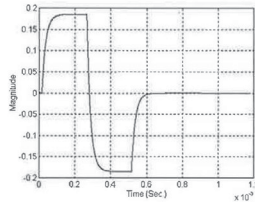
The severe channel intersymbol interference of a twisted-pair loop also can be avoided by signaling at a lower baud rate while letting each baud carry many bits of information. For a 1 mile, 26-gauge twisted-pair loop, the tail of the impulse response is about 30 microseconds. To make the intersymbol interference negligible, the signaling baud interval can be chosen as 100 times that of the tail. This results in a signaling baud rate of 333 Hz.

Data can be carried in this long baud interval with many waveforms of different frequencies. In other words, bits of information are encoded in the frequency domain. This information encoding process can be carried out with the *Inverse Discrete Fourier Transform* (IDFT). Bits of information are recovered in a receiver with the *Discrete Fourier Transform* (DFT). This approach has been called *Discrete MultiTone* (DMT) line code for the use of multiple numbers of carriers. QAM and CAP are called *single-carrier systems*.

A slower baud rate can lead to less intersymbol interference. On the other hand, a slower baud rate requires a higher number of carriers for certain desired transmission throughputs. This in turn results in a bigger IDFT/DFT size. The implementation of a large size IDFT/DFT requires more buffering memories and a longer process delay time.

As an alternative for DMT-based systems, a higher baud rate can be used also, thus lowering the size of the IDFT/DFT requirement. The intersymbol interference at a particular baud rate also can be minimized by the introduction of a guard period where repetitive information is encoded.¹⁹ Figure 8.44 shows the relationship between a long duration baud interval and the tail of the impulse response.

Figure 8.44 Baud Interval Pulse and Residual Tail



Incidentally, the duration of the impulse response tail also can be reduced using the general channel equalization technique. The idea is to use a linear channel equalizer in the receiver front end of a DMT system to minimize the MMSE between the linear equalizer output and an unknown channel with a desired impulse response tail. The solution of this DMT time-domain equalizer has been examined by many studies^{20,21} and for other receiver structures.^{22,23}

The DMT time-domain equalizer solution also can be derived in the discrete time domain. We assume that the equalized unknown channel is the convolution of the channel and the equalizer and can be expressed as

$$h_{\text{equ}}(k) = h(k) * Q(k)$$

To constrain the length of the equalized unknown channel, for an example of an original channel length of 5, an equalizer length of 5, and an equalized channel length of 3, we need to have

$$\begin{bmatrix} h(1) & 0 & 0 & 0 & 0 \\ h(2) & h(1) & 0 & 0 & 0 \\ h(3) & h(2) & h(1) & 0 & 0 \\ h(4) & h(3) & h(2) & h(1) & 0 \\ h(5) & h(4) & h(3) & h(2) & h(1) \\ 0 & h(5) & h(4) & h(3) & h(2) \\ 0 & 0 & h(5) & h(4) & h(3) \\ 0 & 0 & 0 & h(5) & h(4) \\ 0 & 0 & 0 & 0 & h(5) \end{bmatrix} \begin{bmatrix} q_1 \\ q_2 \\ q_3 \\ q_4 \\ q_5 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ h_{\text{equ}}(1) \\ h_{\text{equ}}(2) \\ h_{\text{equ}}(3) \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix}$$

In general, the equalized unknown channel can have an infinite number of solutions. The optimal equalizer solution is called the target channel, which produces the MMSE in conjunction with the equalized channel. Figure 8.45 illustrates the mean square error between the output of a DMT time domain linear channel equalizer and the unknown target channel. The size of the target channel of a DMT system is usually defined by the guard period of the DMT system, and that is normally part of the design parameters.

Figure 8.45 Equalizer and Target Channel

