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Microcoded Modem Transmitters

Abstract: This paper describes various microcoded designs for modem transmitters. The digital echo modulation technique, originally introduced by J-M. Pierret, is applied to cover the case of a fully digital universal modem. The capabilities of several microcoded modem designs are presented and their limitations are discussed.

Introduction

With the present trend of semiconductor technology towards large scale integration (LSI), it becomes economically attractive to implement digitally most of the modem functions.

The paper by A. Croisier and J-M. Pierret [1] has described a new approach called *digital echo modulation* which permits the realization, by digital means, of various filtering and modulation functions in a modem transmitter. This approach is based upon the use of a signal element that is the impulse response of a generalized Nyquist filter.

More recently, the authors have published a paper [2] that extends this concept to cover most of the cases of digital data transmission. Digital echo modulation is a technique in which the line signal in a modem transmitter is synthesized by generating signal elements in a time sequence. References [1] and [2] showed that most of the practical cases of linear modulation are covered by digital echo modulation and that a modem transmitter is characterized by two main parameters: a finite set of signal elements, which characterizes the amplitude spectrum of the line signal, and the algorithm relating the assignment of signal elements to input data, which characterizes the modulation.

Under these conditions, the various digital echo modem transmitters differ in the way the signal elements are stored and in the way the modulation algorithms are implemented.

In the first practical applications, namely the IBM 3978 modem family and the 7200 bps IBM 3875 modem, the conventional transmitting modulators and filters were eliminated and replaced by an early implementation of the digital echo modulation design. In this configuration the signal elements are stored as analog taps and the modulation algorithm is hardwired.

The purpose of the present paper is to describe an all-digital implementation of digital echo modulation. In this approach, the signal elements are stored in a digital memory as pulse code modulation (PCM) samples or as delta modulation samples. We shall show that, by properly implementing the modulation algorithm, it is feasible to design an all-digital multispeed, multifunction modem transmitter that leads to an efficient implementation in large scale integration.

We first briefly review the basic concepts of digital echo modulation. Application to a microcoded modem transmitter is then discussed. Finally, the optimization of digital signal elements is treated and practical results are given.

Digital echo modulation

Most high-performance data transmission schemes are synchronous and use linear modulation techniques. As pointed out previously [1-3], the linear filtering and modulation functions performed in a modem transmitter can be viewed as being produced by a set of elementary signal generators. The result is a set of elementary signals called signal elements.

We consider the elementary modem transmitter shown in Figure 1(a). In such a modem, where the signaling interval is T, the input data is a sequence of impulses of amplitude a_i at sampling time iT. After lowpass filtering with a filter $G(\omega)$, this signal is product modulated with a carrier of angular frequency ω_c and phase φ. The resulting amplitude-modulated suppressed carrier (AMSC) signal is then bandpass filtered with filter $H(\omega)$. This line signal $e(t)$ represents either a double-sideband or a vestigial-sideband modulated signal, depending upon the particular characteristics of the postmodulation filter.

If the impulse responses of $G(\omega)$ and $H(\omega)$ are assumed to be $g(t)$ and $h(t)$, respectively, $e(t)$ can be expressed as

$$e(t) = \left\{ \left[\sum_{i=-\infty}^{i=+\infty} a_i g_i(t - iT) \right] \cos\,(\omega_c t + \varphi) \right\} * h(t)\,, \quad (1)$$

where $*$ is the symbol for the convolution product. In conventional approaches the baseband signal $\sum_{i=-\infty}^{i=+\infty} a_i g_i(t-iT)$ is obtained from a continuous analog filtering of impulses a_i and the modulator and bandpass filter are implemented with analog circuits.

It should be noted, however, that Eq. (1) can be rewritten as follows [1, 2]:

$$e(t) = \sum_{i=-\infty}^{+\infty} a_i e_i(t - iT)\,, \quad (2)$$

with $e_i(t) = [g(t) \cos\,(\omega_c t + \varphi + \omega_c iT)] * h(t)$.

It is clear from (2) that the modem transmitter can be viewed as a signal element generator in which a particular signal element $e_i(t)$ is associated with each sampling time iT. It can be seen that the signal elements are dependent upon the sampling time iT. This relation can be viewed as the condition of carrier phase continuity that must be insured when the transmitter in Fig. 1(a) is replaced by a signal element generator.

This result has been extended to cover all cases of linear modulation, i.e., all cases in which the modulated signal is linearly related to the baseband signal. These cases include amplitude modulation, vestigial sideband modulation, phase shift keying, and linear frequency shift keying modulation, as defined by Sunde [4].

A universal modem transmitter is shown in Fig 1(a). In this modem, multiphase, multiamplitude modulated signals can be generated by assigning the appropriate weightings to amplitude coefficients a_{i1} through a_{iN}. These coefficients activate the N elementary transmitters, each of them corresponding to a discrete value φ_1 to φ_N of the carrier phase. The tone generators are used to generate the tones required either in the case of linear frequency shift keying modulation or the case when side tones must be generated in order to allow for carrier and clock recovery in the receiver. Under these conditions, the line signal for any linear modulation can be expressed as

$$e(t) = \sum^{+\infty} \sum_{j=1}^{+N} a_{i,j} e_{i,j}(t - iT) + \sum_{e=1}^{L} d_e \cos\,(\Omega_e t + \alpha_e)\,, \quad (3)$$

where

$$e_{i,j}(t) = [g(t) \cos\,(\omega_c t + \varphi_j + \omega_c iT)] * h(t)\,, \quad (4)$$

in which φ_j is the carrier phase at sampling time iT and ω_c the angular carrier frequency.

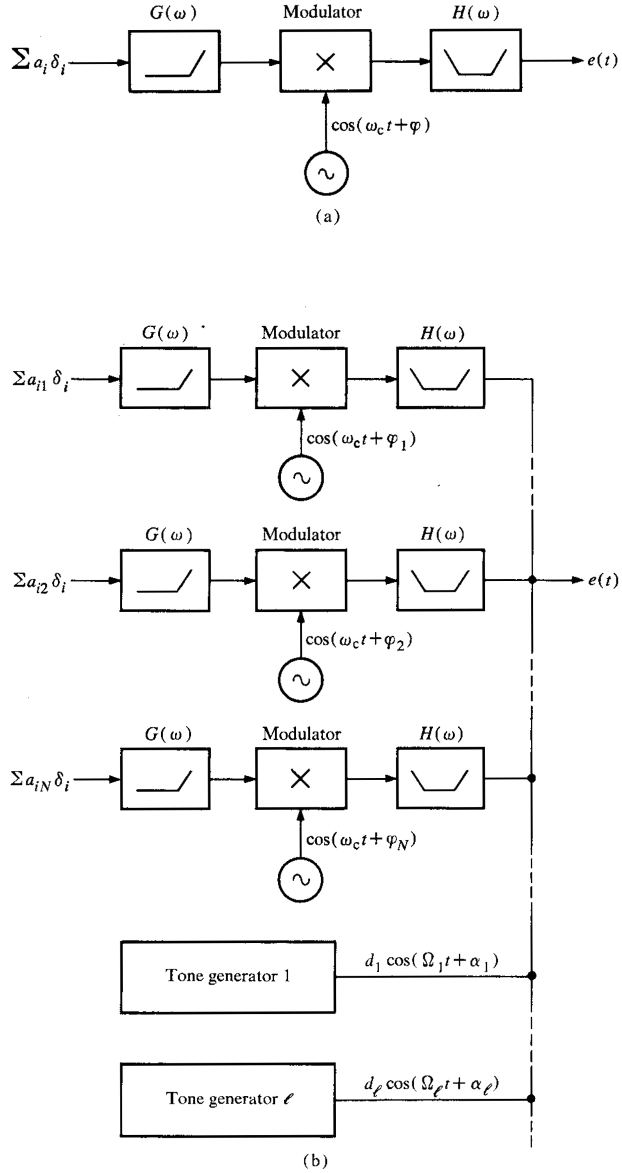


Figure 1 (a) Elementary modem transmitter. (b) Universal modem transmitter.

With $E_{i,j}(\omega)$ assumed to be the Fourier transform of $e_{i,j}(t)$, the generalized modem transmitter can be viewed as a signal element generator, as shown in Fig. 2. This generator is designed to send N sets of signal elements. At each sampling time, one or several among the N possible signal elements are fetched from a signal element store as a function of the input data. The tones are generated continuously as a function of the desired tone pattern.

At this point it is interesting to define more precisely the relations between double-sideband (DSB) and

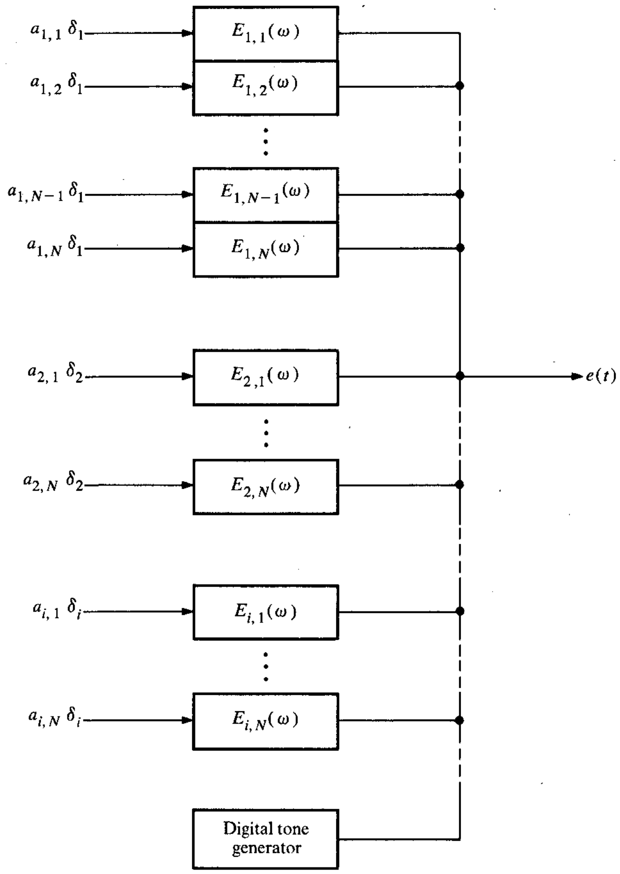


Figure 2 Universal modem transmitter viewed as a signal element generator.

vestigial-sideband (VSB) modulation. Double-sideband modulation can be considered as being generated by a sum of signal elements, provided that phase continuity of the carrier located at midband is assured. Because vestigial-sideband modulation can be derived by filtering double-sideband modulation, it can also be considered as generated by a sum of signal elements. However, in the case of vestigial-sideband modulation, the carrier is located at the edge of the signal spectrum and therefore differs from that of the signal elements. We show how carrier phase continuity can be maintained in these conditions.

Let us assume that we wish to generate DSB and VSB signals with the same line spectrum.

We define ω_c as the angular carrier frequency of DSB modulation, $1/T_D$ as the signaling rate of DSB modulation, ω_0 as the angular carrier frequency of VSB modulation, and $1/T_V$ as the signaling rate of VSB modulation.

We next show that relations (3) and (4) can be used for both types of modulation except that φ_j is always zero and $\omega_c iT$ becomes $\omega_0 iT_V$ for VSB.

Noting that $\omega_0 = \omega_c \pm 2\ \pi/4T_V$ and $T_D = 2T_V$, the carrier phase continuity condition, given by $\omega_c T_D$ in the DSB case becomes, for the VSB case.

$$\omega_0 iT_V = \omega_c iT_V \pm i\pi/2 = \omega_c iT_D/2 \pm i\pi/2.$$

VSB modulation can then be viewed as a particular case of DSB multiphase modulation in which the carrier phase shift $\varphi_j + \omega_c iT_D$ is replaced by $i\pi/2 \pm \omega_c iT_D/2$.

It has been shown [2] that the number of signal element is finite provided that

$$\omega_c T/2\pi = P/Q, \tag{5}$$

where P and Q are integers.

In this case, the signal elements corresponding to a given data value are a periodic function of the sampling time iT, as shown by the basic definition formula. Under these conditions, if N is the number of equally spaced phases and M the number of different amplitudes, the number of different signal elements reduces to the product of M by the least common multiple between N and Q.

In practice, the modem transmitter stores and generates signal elements of finite duration derived from the "theoretical" signal element by a truncating window. It has been shown [1, 2, 5] that a good approximation of the ideal spectrum can be achieved with signal element durations limited to a few signaling intervals.

As pointed out above, the various linear modulation schemes can be decomposed into a set of signal elements. Under these conditions, a modem transmitter can be viewed as a signal element generator. In this approach, the conventional analog modulators and filters in a modem transmitter are replaced by logic circuitry and a signal element memory. The wave shapes of the stored signal elements characterize the amplitude spectrum of the line signal, and the modulation is defined by the algorithm relating the input data to the fetched signal elements.

In the first applications of digital echo modulation to practical modem transmitters, the signal elements were stored as analog taps in a shift-register-like arrangement and the modulation algorithm was hardwired with logic circuits. This arrangement is shown schematically in Fig. 3. The present approach is well suited for implementing relatively simple modem transmitters operating at 2400 bps or 4800 bps. However, when it is desired to build multispeed, multifunction modems, the signal element storage requirements increase significantly and it becomes more attractive to store the signal elements as PCM or delta modulation samples in a digital memory. We shall show in the following section how this approach can be implemented to take full advantage of large scale integration for designing such modems.

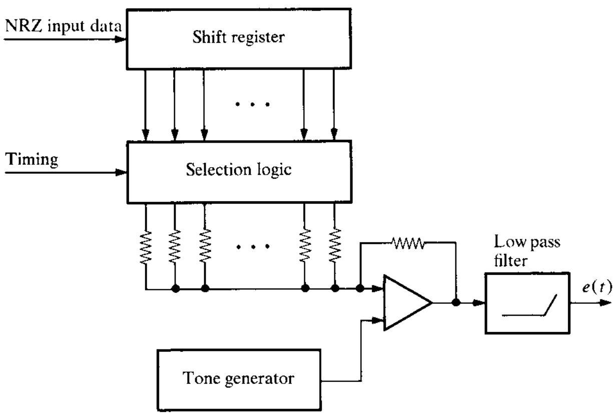


Figure 3 PAM digital echo modem transmitter.

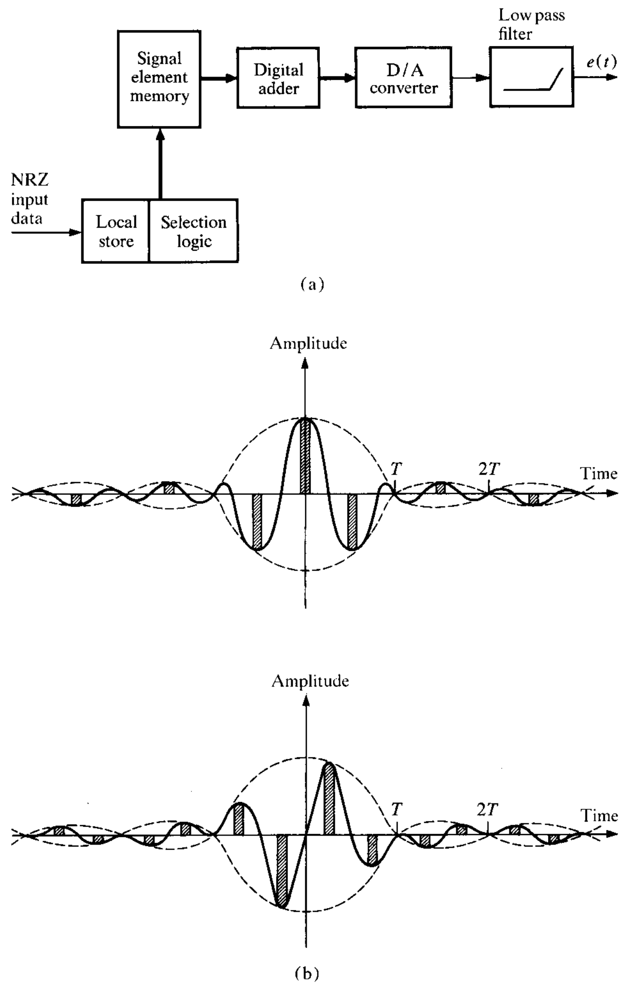


Figure 4 (a) PCM or delta digital echo modem transmitter. (b) Signal element generation and encoding of "in phase" and quadrature signal elements. (See text).

Microcoded transmitter

We have seen in the preceding section that, provided some relation between data rate and carrier frequency is maintained, the line signal can be synthesized from a finite set of time-limited signal elements. Under these conditions, the modem transmitter can be viewed as a signal element generator, as shown in Fig. 4(a).

From a practical viewpoint, it is highly desirable to store the various signal elements in a digital memory. This can be done easily since the signal elements are time domain signals which have a band-limited spectrum. Under these conditions, the analog signal elements can be replaced by their sampled representation, i.e., by a discrete number of impulses (PAM representation). The amplitude of those impulses can in turn be binary encoded as PCM samples so that the signal elements are represented by a set of bits and can therefore be stored in a digital memory.

The operation of the microcoded modem transmitter in Fig. 4(a) can be understood by considering the simple example illustrated in Fig. 4(b). In the example, we consider the case of a four-phase modem with $\omega_c T = 2\pi$. This case would correspond, for example, to that of a 3200 bps modem operating with a 1600 Hz carrier frequency. Such a modem requires only the two signal elements shown in Fig. 4(b) and the two opposite values. These signal elements are stored in the digital memory as a set of PCM samples, with, for instance, four samples per signal interval. At every new signaling time, that is to say, every 1/1600 s, the selection logic fetches two new bits of input data and stores them temporarily. The selection logic selects the signal element which corresponds to the dibit configuration and fetches at regular intervals the various samples that correspond to this signal element. The process goes on until the signal element dies away. At this point, the old dibit is replaced by a new dibit in the local store and the process repeats itself.

Usually, the signal element duration extends over several signaling times so that several dibits must be stored temporarily and several signal elements must be generated simultaneously. If we assume, for instance, that the roll-off of the spectrum is such that each signal element extends over four signaling intervals, at any given time, the line signal is the sum of four signal elements and the local store must have a capacity of four dibits.

Under these conditions, each line-signal sample can be constructed by fetching in an interleaved mode the four PCM samples corresponding to the four dibits stored in the local store and accumulating them in a digital adder. The selection logic insures that the succession of signal element samples is fetched in the proper order. The output of the digital accumulator is a sequence of PCM words. These words are converted into analog impulses by a digital-to-analog converter. The line signal $e(t)$ is then reconstructed by filtering those impulses with a low-pass filter, so as to eliminate upper sidelobes.

The digital modem transmitter can thus be viewed as being composed essentially of a signal-element store, a local read-write data store, and logic circuitry. These three basic components can be arranged in a variety of ways to fit a particular modem design. With the advent of LSI, a better approach would be to have a universal or near-universal design that could be "personalized" to fit particular modem requirements by specifying read-only storage patterns and by using strap and switch arrangements.

This view leads to a more precise definition of the basic parameters that characterize a modem transmitter. The two most important are the data signaling rate and the line signaling rate. In practice, there is an integer relationship between the line data rate and the signaling rate in which the number of bits per baud can vary from 1 to 6 (1 would correspond to a two-phase modem and 6 would correspond to two eight-level orthogonal channels). These two parameters can be specified with two numbers. In practice, we will specify those two parameters with three words, W1, W2, W3. The word W1 characterizes the master oscillator frequency. W2 specifies the bit rate with respect to W1, and W3 specifies the data signaling rate with respect to W1. More precisely, W3 defines the number of encoded bits at each signaling time and the number of different phases and amplitudes which characterizes the modulation process.

We have seen in the second section that successive signal elements must satisfy the condition of carrier phase continuity with $\omega_c T/2\pi = P/Q$. This condition will be specified with a word W4 which will be the coded value of $\omega_c T$.

The number of words in the local store is proportional to the signal element duration expressed in number of line-signaling intervals, and its word size in number of bits determines the maximum number of signal elements used in the modem. In practice, the modems with the sharpest roll-offs will not require signal element durations longer than eight line-signaling intervals nor more than 64 signal elements. This means that the needs will be covered with a local store of 48 bits (eight 6-bit words). This local store can conveniently be implemented with shift registers, as will be shown later.

The content of the signal element store defines the spectrum of the line signal. We will show later on that a single store configuration can cover all practical needs.

Aside from the signal element store personalization, we have seen that the modem transmitter could be characterized by only four words, W1 to W4. A practical modem transmitter that can be personalized easily to fit particular speed and modulation requirements is shown in Fig. 5. In this approach the design rationale has been to devise a universal modem transmitter in which all timing pulses are derived from a single master oscillator and in which most of the wired logic has been replaced by standard arithmetic units and memories. Moreover, the design has been optimized so as to reduce the memory size as much as possible. This transmitter can be programmed to operate in a variety of modes and at various speeds but is particularly well adapted for operation in the 1200 bps to 10,800 bps range. Some practical modem transmitters that can be built with the microcoded implementation are listed in Tables 1 and 2.

In the particular implementation of Fig. 5, the signal-element sampling frequency F_e is generated from the oscillator frequency by accumulating an eight-bit word W1 at each oscillator period. Accumulator overflows determine a multiple (10 or 12) of the sampling frequency F_e. A set of timing circuits, under control of a word W2, generates the transmitter clock frequency and 10 equally spaced timing pulses T_1 to T_{10}.

The line-signal spectrum is characterized by the content of a signal element read-only store, organized as 2048 words of 8 bits. The various signal elements are selected through address lines 1 to 4, and the various samples are addressed through control lines 6 to 11.

In practice, the number of signal element and sample selection lines varies, depending upon the particular transmitter to be programmed. Memory utilization is optimized by allowing address line 5 to be used under control of W3 either as an additional signal element selection line or as an additional sample selection line.

Counters C_2, C_3, C_4 determine the number of PCM samples per signaling interval (6 or 12) and generate the signaling frequency $1/T$, which can be set at $F_e/6$ or $F_e/12$ under control of W2. At each period of the sampling frequency F_e, the divide-by-eight counter C_1 selects in sequence the eight signal element samples corresponding to a sample of the line signal. These samples are accumulated so as to synthesize a digital sample of the line signal. The analog line signal is derived from the sample through a digital-to-analog converter followed by a low-pass filter aimed at eliminating upper sidelobes.

We have seen that up to five ROM addressing bits are devoted to signal element selection and that the signal element duration is set to eight signaling intervals T. This approach allows the use of up to 32 different signal elements. This range can be extended to up to 64 signal elements because signal elements which differ only by a phase difference of π in their carrier frequency have the same amplitude and opposite signs. Therefore a sixth selection bit is used to complement the output of the

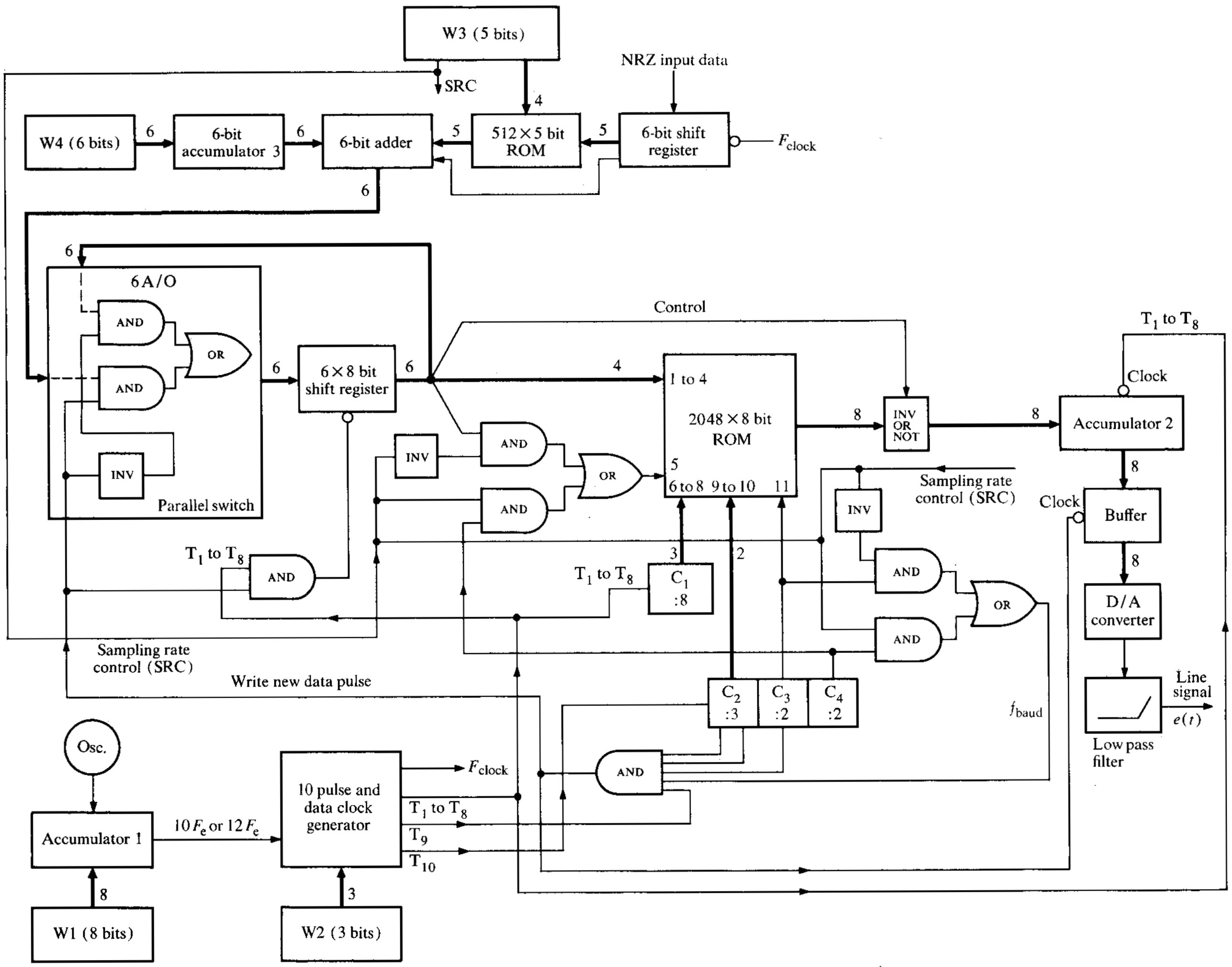


Figure 5 PCM microcoded transmitter. Multiline circuits are indicated by numerals.

signal element store so as to double the number of usable signal elements at little additional cost.

Under these conditions, at each signaling period, six new data bits must be entered into the transmitter. This group of bits replaces the oldest group of six bits in a local shift register store of eight six-bit words. This group of bits can be extracted from a six-bit shift register operating at the clock frequency. A read-only store of 512 words, five bits is arranged in such a way as to select the number of bits per line signaling interval, which can vary from one to six, and to rearrange those bits in such a way that the rightmost bits characterize the various carrier phases, while the leftmost bits characterize the various amplitudes. The phase bits b_1 to b_n are arranged in decreasing order such that the carrier phase can be written as

$$\theta_1 = b_1\pi + b_2\pi/2 + \cdots b_n\pi/2^n. \qquad (6)$$

It should be noted that the 512×5 bit read-only memory controlled by the five-bit word W3 can be replaced by a smaller read-only memory of 32×5 bit words. In this case the five-bit word W3 is replaced by a 160-bit specification of this new read-only memory. Therefore this approach allows a slight cost reduction at the expense of reducing somewhat the flexibility in mode of operation.

We have seen in the second section, Eq. (2), that at each signaling interval, the signal element carrier phase is shifted by $\omega_c iT$. This means that before being transferred to local store for signal element selection, the

Table 1 PCM microcoded modem. Examples of VSB modem characteristics.

Signaling rate $1/T$ (baud)	*Frequencies for 40 dB out-of-band attenuation*		*Line spectrum center frequency*	*Upper Nyquist limit*	*Number of phases*	*Number of amplitudes*	*Speed (bps)*	*Number of signal elements*	*Tone frequencies*		$\frac{\omega_c T}{2\pi}$
1200	2175	1425	1800	2100	2	1	1200	4	2400	1200	7/4
						2	2400	8			
						4	3600	16			
1800	2212.5	1287.5	1650	2100	2	1	1800	6	2600	2600	7/6
						2	3600	12			
						4	5400	24			
2400	2550	1050	1800	2400	2	1	2400	2	2800	800	1
						2	4800	4			
						4	7200	8			
3200	2600	600	1600	2400	2	1	3200	4	2900	400	3/4
						2	6400	8			
						4	9600	16			
3600	2700	450	1575	2475	2	1	3600	16(6)	2895	375	11/16
	2625	375	(1500)	(2400)					(2810)	(350)	
						2	7200	32(12)			
						4	10 800	64(24)			

group of data bits must be updated for carrier phase continuity by adding $\omega_c iT$ to θ_1 so that the new signal element becomes

$$\theta_2 = \theta_1 + \omega_c iT$$
$$= \theta_1 + \sum \mathrm{W4},$$

with $\mathrm{W4} = \omega_c T = 2\pi f_c T$.

The running carrier phase is computed by accumulating W4 at each signaling period. The resulting number is added to the data phase bits and the resulting updated phase bits are transferred to the local store along with the amplitude bits.

Signal element selection is performed under control of the eight-word-six-bit shift register local store. Each group of six bits is stored for the duration of eight signaling intervals and is replaced by a new group at the end of this period, at time T_g, under the control of logic circuits.

The transmitter design shown in Fig. 5 can be modified in a number of ways to fit particular requirements in circuit technology or in modem specifications: for instance, the various adders can be either replaced by a single adder shared between the various units or implemented as serial adders. This modification is generally feasible because the adder speed requirements are relatively modest: no more than 100,000 to 200,000 eight-bit additions per second for ACC 1 and ACC 2.

The design shown in Fig. 5 has been deliberately restricted to the signal processing part of the modem transmitters. Other functions such as initialization and interface control, scrambling, Gray code converter, etc., have not been included because in some cases they are part of the terminal rather than part of the modem. Moreover, the low instruction rate required for processing those functions makes them ideally suited for treatment in a conventional computer or microprocessor and therefore is beyond the scope of the present paper.

It should be noted that if the two ROM's are replaced by two RAM's, and if loading circuitry is added in order to update the memory content and words W1, W2, W3 and W4, the modem transmitter speed and mode of operation can be completely remote controlled from a terminal or a computer.

Microcoded modem transmitter capabilities

The microcoded modem transmitter described in the preceding section can be programmed to operate in a variety of modes and speeds. In this section we attempt to determine the various modem designs that can be implemented with the present approach.

The sampling frequency, clock, and signaling frequency generation circuits have been designed with sufficient flexibility to cover all practical cases of voice-grade line modems while maintaining the sampling frequency within sufficiently tight limits to keep a reasonable balance between the output low-pass filter requirements and signal-element storage requirements. The sampling frequency can be either six times or 12 times the signaling frequency, and the clock frequency can be set at any integer multiple between one and six of the signaling frequency. This insures a minimum sampling frequency

Table 2 PCM microcoded modem. Examples of multiphase, multiamplitude modem characteristics.

Modulation type	*Signaling rate, 1/T (baud)*	*Frequencies for 40 dB out-of-band attenuation*		*Line spectrum center frequency*	*Upper Nyquist limit*	*Number of phases*	*Number of amplitudes*	*Speed (bps)*	*Number of signal elements*	$\frac{\omega_c T}{2\pi}$
Phase	1200	2550	1050	1800	2400	2	1	1200	2	3/2
Phase	1200	2550	1050	1800	2400	4	1	2400	4	
Phase + amplitude	1200	2550	1050	1800	2400	4	2	3600	8	
Phase	1200	2550	1050	1800	2400	8	1	3600	8	
Phase	1600	2800	800	1800	2600	8	1	4800	8	9/8
		(2600)	(600)	(1600)	(2400)					9/8
Phase + amplitude						8	2	6400	16	
						8	4	8000	32	
Phase	1800	2925	675	1800	2700	2	1	1800	2	1
Phase						4	1	3600	4	
Phase + amplitude						4	2	5400	8	
Phase + amplitude						8	2	7200	16	
Phase + amplitude	1920	2880	480	1680	2640	8	4	9600	32	7/8
		(3000)	(600)	(1800)	(2760)				(64)	
Phase + amplitude	2400	3300	300	1800	3000	4	1	4800	4	3/4
						8	1	7200	8	
						8	2	9600	16	

of 9600 Hz for a 800-baud modem and a maximum sampling frequency of 14,400 Hz for a 2400-baud modem. The minimum sampling frequency of 9600 Hz is sufficiently high to allow easy elimination of the upper line-signal spectrum sidelobes by a low-cost, low-pass filter. The signal-element storage requirements are directly related to the sampling frequency and correspond to 64 or 128 eight-bit words per signal element, depending upon the ratio between sampling and signaling frequencies.

One of the main limitations of the proposed design concerns the ratio between carrier frequency and signaling frequency. These limitations have already been discussed in the second section and can be translated into the fact that unfavorable ratios between carrier and signaling frequencies could require a very large number of different signal elements. It can be shown, however, that a large flexibility is retained with a maximum storage capability of 64 signal elements.

If we take, for example, the case of an N-phase modulation with equally spaced phases, and if R is the number of stored signal elements, the condition (5) of carrier phase continuity becomes

$$f_c T = i/R \qquad \text{with } Nn = R, \quad n = \text{integer}. \tag{7}$$

In the case of a four-phase 2400 bps modem, $1/T =$ 1200 baud and $f_c = i \times 1200/64 = i \times 18.75$ Hz. The carrier frequency can therefore be set at any integer multiple of 18.75 Hz. Under these conditions, it can be seen that practically all real-life cases of phase modulation can be covered with the proposed approach.

In the case of a mixture of amplitude and phase modulation, some signal elements must be reserved for amplitude modulation, so that the flexibility in choosing the carrier frequency is somewhat reduced. For instance, in the case of a 4800-bps modem operating at 1200 bauds with two four-level orthogonal channels, the number of phase bits reduces to four. Under these conditions, the carrier frequency can be set at any integer multiple of 75 Hz.

Some practical modem transmitters that can be built with the microcoded implementation are listed in Tables 1 and 2. Table 1 lists various VSB modem transmitters operating at speeds between 1200 and 10,800 bps with various combinations of bandwidth, carrier frequency and data rate. Table 2 lists various phase- and amplitude-modulated modems. It should be noted that, in this table, the 2400 bps four-phase modem corresponds to CCITT recommendation V26, alternative A, while one of the 4800 bps eight-phase modems conforms with several CCITT contributions [6, 7] presently considered for standardization.

Modem transmitters with interleaved signal elements

In the preceding section, we have described a flexible general-purpose transmitter that can readily be implemented with LSI and in which analog circuitry is reduced to a digital-to-analog converter. We show in the following sections that the digital hardware count can be significantly reduced by generating the signal elements in an interleaved mode and the digital-to-analog converter can be greatly simplified by encoding the signal elements as delta modulation samples instead of PCM samples. These two points are discussed in the particular cases of a four-phase and an eight-phase modem transmitter.

• *Four-phase transmitter*

Let us consider a four-phase transmitter operating at 2400 bps with carrier frequency at 1800 Hz and a signal element extending over four signaling intervals. In such a modem, the signaling rate is 1200 baud, the Nyquist bandwidth is 1200 Hz and the line signal spectrum bandwidth for a 40 dB out-of-band attenuation is 1800 Hz.

We have seen above that for signal elements extending over U signaling intervals, the current line signal sample was constructed by adding the U weighted samples corresponding to U overlapping signal elements. In practice, the U signal elements are fetched in sequence from the ROM and added in an accumulator so that the PCM decoder operates at a low frequency of about 10,000 to 20,000 samples per second.

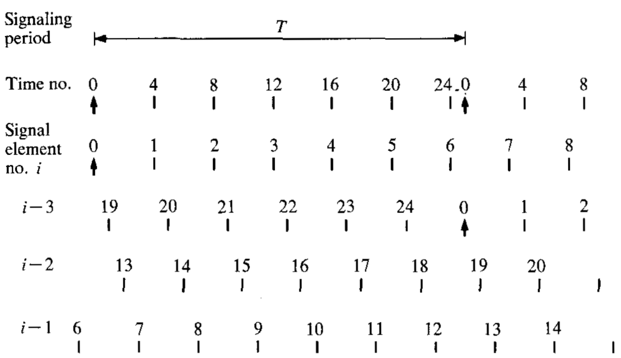
The accumulator can be eliminated by feeding the signal element samples directly to the PCM decoder. With this approach, actual addition is replaced by an interleaving process and the PCM decoder operates at U times the previous sampling rate, which is still a moderate speed. Care must be exercised, however, in order to insure that the various signal element samples are fetched at times that correspond to the encoding time. If this were not the case, distortion would be introduced by the time jitter caused by the interleaving process.

The encoding can be made to be compatible with the interleaving mechanism by choosing a signal element sampling frequency F_e such that

$$F_e = (K + 1/U)/T \qquad (8)$$

instead of taking $F_e = K/T$, as in the conventional case.

Figure 6 Transmitter timing. Four-phase interleaved PCM coding.



Under these conditions, and in our example where $K = 6$, $U = 4$ and $1/T = 1200$ Hz, the signal element sampling frequency $F_e = (6 + 1/4)/T = 7500$ Hz, and there are 25 samples per signal element. The interleaving mechanism is shown in Fig. 6 and the four-phase transmitter block diagram is given in Fig. 7. In this modem, the apparent output sampling rate is $F_K = 4(6 + 1/4)T = 25/T$, and all clock frequencies are derived from a master oscillator operating at $F_{oc} = 50/T$.

The 2400-Hz data clock frequency is derived from F_{oc} by a divide-by-25 counter. The clock frequency F_K, derived from F_{oc} by a divide-by-2 counter, serves to drive in cascade a divide-by-4, followed by a divide-by-8 ROM address counter. At each sampling time, one of the 25 signal element samples is addressed by these counters.

At address 24, a separate ROM bit line (write new data pulse) corresponding to sample number 6 fetches a 1, indicating the beginning of a new signaling interval. The various counters are then reset and the oldest dibit is replaced by a new one. The condition of carrier phase continuity is given by $\omega_c T = 2\pi\ 1800/1200 = 3\pi$. This condition means that a phase π must be added to the current signal phase at each signaling interval. In practice, this condition is met by inverting the dibit sign at every other signaling time thanks to a divide-by-2 counter operated by the "write new data pulse" line.

The other parts of the modem transmitter operate in the same way as in the preceding section. It should be noted that the ROM size reduces to only two separate groups of 25 words of eight bits each, or a total of 400 bits.

• *Eight-phase transmitter*

In the previous sections, we have assumed that the signal elements were encoded as PCM samples. This meant that a PCM decoder followed by a low-pass filter had to be used to reconstitute the analog line signal. An attractive alternative is to encode the signal elements by using the delta modulation technique [8, 9]. With this approach, and provided an interleaving scheme is used, the analog output circuits can be reduced to a simple integrator.

In the following, we will consider such a design in the case of an eight-phase modem operating at 4800 bps with a 1800-Hz carrier frequency and a signal element extending over eight signaling intervals. In such a modem, the signaling rate is 1600 baud, the Nyquist bandwidth is 1600 Hz, and the line signal spectrum bandwidth for a 36 dB out-of-band attenuation is 2000 Hz.

In order to insure proper interleaving, the signal element delta sampling frequency F_e must be chosen such that $F_e = (K + 1/U)/T$. Taking $U = 8$ and $K = 61$, F_e is equal to 97 800 Hz, a frequency large enough to insure a good signal-to-noise ratio at the transmitter out-

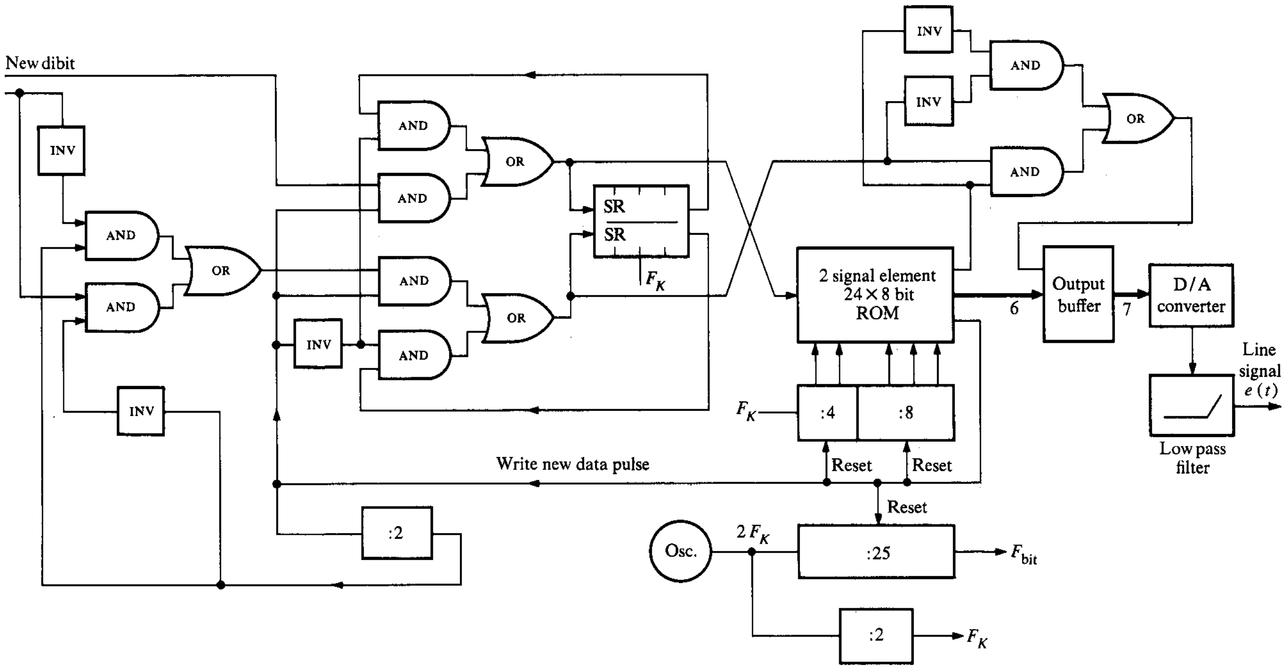


Figure 7 Four-phase transmitter, interleaved PCM coding.

put. The interleaved output frequency is $F_K = 8F_e = 489/T$ or $F_K = 3 \times 163/T$. The value of K has been chosen such that the data clock frequency can be derived by dividing F_K by 163.

The interleaving scheme is shown in Fig. 8 and the delta encoded modem transmitter is shown in Fig. 9. Since this transmitter bears close resemblance to previous designs, we will discuss only a few particular points.

The number of bit samples per signal element is 489. The read only memory (ROM) address counters are reset on condition 488. This condition, which corresponds to a new signaling interval, causes a new group of three data bits to replace the oldest group in the shift register store.

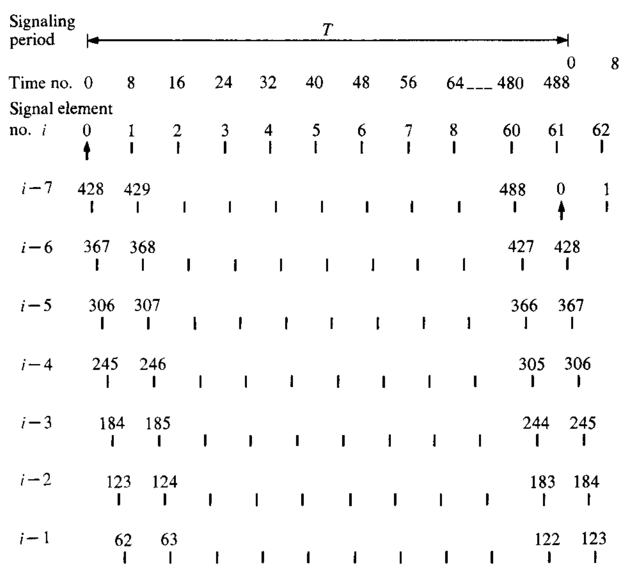
The condition of carrier-phase continuity is given by

$$\omega_c T = 2\pi\ 1800/1600 = 2\pi + \pi/4.$$

This condition means that a phase $\pi/4$ must be added to the current signal phase at each signaling interval. This condition is met by accumulating 001 into divide-by-eight counter C_1 for each new group of three input bits and adding the resulting current carrier phase to the input bits into a modulo 8 serial adder.

The ROM size is now 1956 bits, a somewhat larger figure than in the previous case, because a larger number of signal elements is used and because delta encoding is less efficient than PCM encoding. It should be noted, however, that the overall digital design remains relatively simple in spite of the significant cost reduction afforded in the digital-to-analog conversion by the delta coding approach.

Figure 8 Transmitter timing. Eight-phase delta coding.



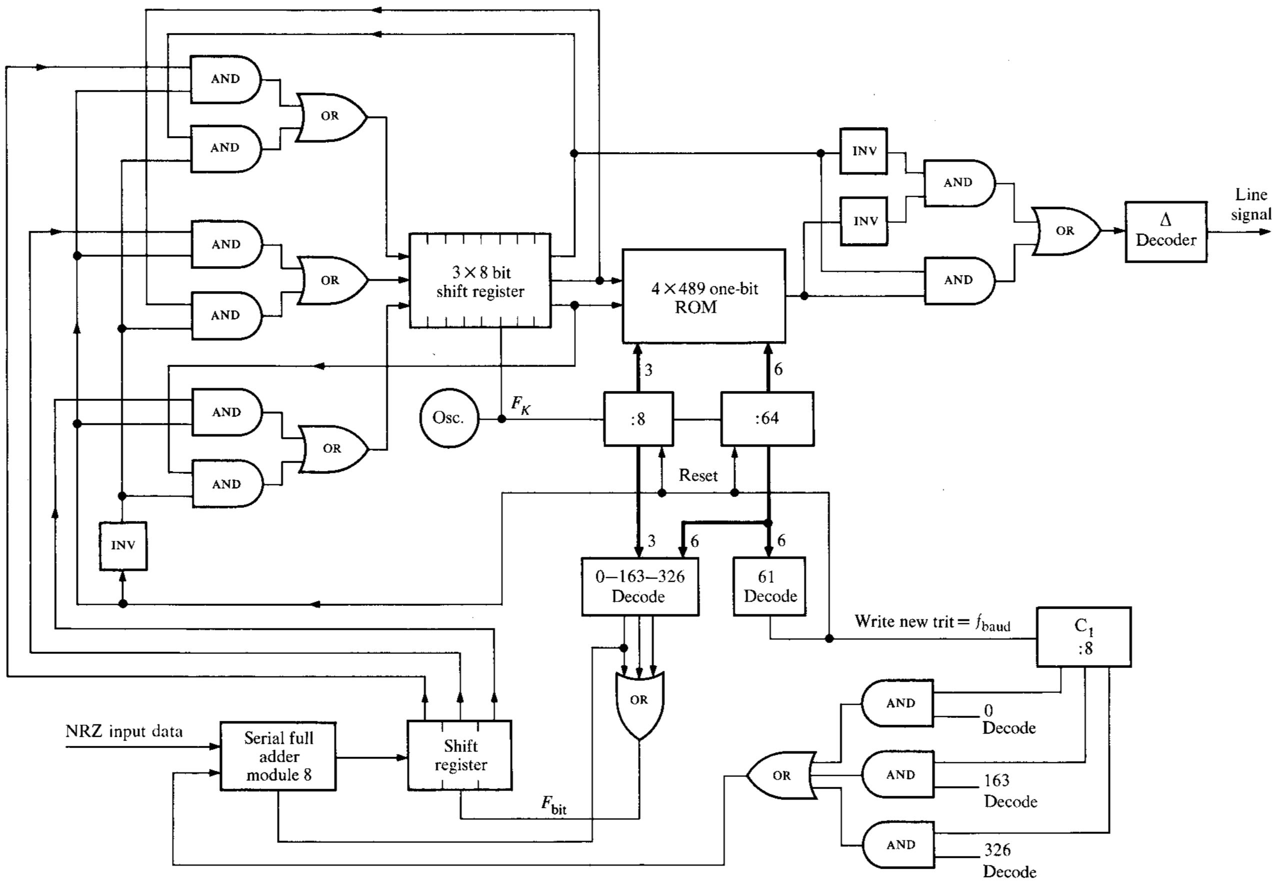


Figure 9 Eight-phase interleaved transmitter with delta-coded signal element.

Design of signal elements

We have seen in the preceding sections how the signal elements should be specified and used for generating a given line signal. These "theoretical" signal elements have an infinite duration and must therefore in practice be truncated and PCM or delta encoded in order to be used in a modem transmitter. Signal element truncation and encoding play a major role in overall modem performance. In the following sections we will briefly review those two methods.

• *Signal element truncation*

Signal elements can be viewed as impulse responses of nonrecursive filters. Under these conditions, the well-known truncating techniques are applicable for designing signal elements. In these approaches, the signal element is weighted by a time-limited window $x(t)$. This corresponds in the frequency domain to a convolution between the signal-element spectrum and the window spectrum.

If we take the case of a Hamming window [5] of duration τ, $x(t)$ is defined by

$$\begin{aligned} x(t) &= 0.54 + 0.46 \cos \pi t/T \qquad &&\text{for } |t| \le \tau \\ x(t) &= 0 &&\text{for } |t| > \tau. \end{aligned} \tag{9}$$

This window is such that 99.96% of its energy is in its main lobe ($|\omega| \le 2\pi/\tau$) with sidelobes peak amplitudes being less than 1% of that of the main lobe.

Under these conditions, with an assumed bandwidth of $1/T + 2\Delta f$, the roll-off $2T\Delta f$ obtained by weighting the impulse response of a rectangular filter with a bandwidth $1/T$ and a Hamming window of duration $\tau = UT$ is

$$2T\Delta f = 2T\ 2/UT = 4/U. \tag{10}$$

A similar computation using Dolph-Chebychev's window [6] with 1% peak sidelobe ripple gives

$$2T\Delta f \approx 3.8/U. \tag{11}$$

It can be seen from (10) and (11) that these conventional approaches lead to the use of signal elements of relatively long duration in order to obtain reasonably sharp roll-offs. This is illustrated by the fact that signal elements must extend over eight signaling intervals in order to achieve a roll-off of only 50%.

Reducing signal element duration for a given roll-off is of primary importance in order to reduce the size and cost of the transmitter. This reduction can be achieved in practice by taking advantage of the fact that, in a modem transmitter, high sidelobe rejection is needed only in the immediate vicinity of the main lobe. The rest of the sideband noise is eliminated by the transmitter output filter, the transmission line and the modem receiver input filter. Moreover, the main lobe ripple need not be much better than the residual distortion on the equalized transmission line.

This method leads to the use of a heuristic approach for optimizing signal element truncation. The technique consists in sampling the impulse response of a trapezoidal spectrum that approximates the desired line signal spectrum. The spectrum corresponding to the time-limited sampled impulse response is computed for various slopes of the trapezoid, and the set of samples that allows maximum near-sideband energy rejection is selected.

More precisely, since that the trapezoid is the convolution of two rectangular spectrums, the impulse response can be written

$$e(t) = \frac{\sin 2\pi\Delta ft}{2\pi\Delta ft} \frac{\sin \pi t/T}{\pi t/T} \cos(\omega_c t + \varphi), \tag{12}$$

$1/T$ being the line signaling rate, $2T\Delta f$ the relative roll-off and ω_c the angular center frequency of the main lobe.

In this approach, the truncating window is defined as

$$\begin{aligned} x(t) &= \sin 2\pi\Delta ft/2\pi\Delta ft \qquad \text{for } |t| \leq UT/2, \\ x(t) &= 0 \qquad \text{for } |t| > UT/2. \end{aligned} \tag{13}$$

Figure 10 illustrates the method, giving the computed and measured spectra of a signal element extending over eight signaling intervals. This spectrum corresponds to a signaling rate of 1600 bauds, and eight-bit PCM encoding and a sampling rate of 10,000 Hz. The signal element is coded with 49 eight-bit samples. It can be seen that for a roll-off of 25%, a 40 dB rejection of out-of-band energy is achieved in the frequency bands 300 to 800 Hz and 2800 to 3400 Hz. It should be noted that this figure represents a factor of two improvement over that obtained with the conventional truncating methods. This roll-off compares favorably with the best roll-offs obtained in conventional modems. In the case of modems with a limited number of levels (up to four levels for amplitude modulation or eight phases for phase modulation), the out-of-band rejection requirements can be decreased to 30 to 35 dB instead of 40 dB, thereby allowing even sharper roll-offs with the same signal element duration.

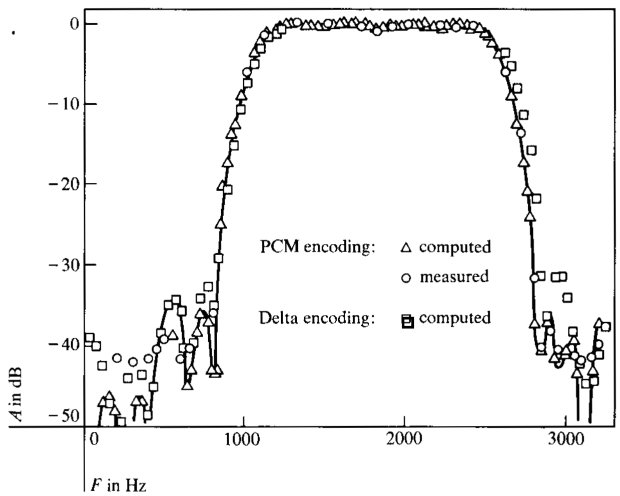


Figure 10 PCM encoding: Computed and measured spectrums. Delta encoding: Computed spectrum.

• *Signal elements encoding*

The digital encoding must be such that quantization noise, roundoff errors and aliasing result in a negligible distortion of the line signal.

In the case of PCM, we have seen that the sampling rate was chosen to be 6 or 12 times the modulation rate so that upper sidebands can be eliminated with simple and noncritical low-pass filters. The number of bits per sample depends upon the required accuracy for the line spectrum. The signal-to-quantization noise ratio is given by the approximate relationship [9]

$$S/N \approx 6V + 2, \tag{14}$$

where S/N is the signal-to-encoding-noise ratio in dB in the Nyquist bandwidth (from zero frequency up to half the sampling rate), and V is the number of bits per sample. It can be seen that a signal-to-noise ratio better than 40 dB is easily obtained with seven or eight bits per sample.

Care must be exercised however in order to minimize round-off errors. In practice, if samples are to be encoded with 2^V levels, one may proceed by determining first the largest sample S_M, normalizing every value with $2^V/S_M$ and rounding those values.

After these operations, the signal elements normally

Table 3 PCM encoding: signal element samples. Exact and quantified values.

Sample number	Exact value	Quantified value	Sample number	Exact value	Quantified value
0–48	0.0039	0	12–3	0.0044	− 1
1–47	−0.0025	0	13–35	−0.0830	−10
2–46	−0.0257	−3	14–34	−0.0761	−10
3–45	−0.0202	−3	15–33	0.1056	13
4–44	0.0135	2	16–32	0.1736	22
5–43	0.0145	2	17–31	0.0245	3
6–42	−0.0035	0	18–30	0.0156	2
7–41	0.0289	4	19–29	0.1768	22
8–40	0.0622	8	20–28	−0.0403	− 5
9–39	0.0026	0	21–27	−0.6062	−76
10–38	−0.0660	−8	22–26	−0.5563	−70
11–37	−0.0306	−4	23–25	0.3865	49
			24	1.0000	126

contain a small dc component, which can be eliminated by modifying the signal elements sample in such a way that their sum is equal to zero.

The signal element computed values and the corresponding quantified values are shown in Table 3. The corresponding spectra are shown in Fig. 10.

In the case of delta encoding, the set of delta bits can be obtained from a simulation of a physically realizable encoder. This will give results conforming with published data. For instance, encoding with double integration at 97,800 Hz, the signal element whose spectrum is shown in Fig. 10 would give a quantization signal-to-noise ratio of 34 dB in the 300 to 3400 Hz band.

A more efficient approach can be used by taking advantage of the fact that the delta encoder is not used "on line". This means that it is not necessary to simulate an actual encoder, but rather to select the particular set of delta bits which gives the best approximation of the desired signal element with the best signal-to-noise ratio. Unfortunately, the number of delta bits is so large (493 in our example) as to preclude a test of all possible sets. Because noise and signal are strongly correlated but are not linearly related, there is no convergent iterative algorithm for improving the encoding.

A significant improvement of signal-to-noise ratio has been obtained by using the following heuristic approach. In a first step, the signal element $e_{i0}(t)$ is encoded into a simulated delta encoder. This gives a first set of bits. In a second step, a part of the difference $\epsilon_0(t)$ between the original signal and the output of the decoder is added to $e_{i0}(t)$. The resulting signal $e_{i1}(t) = e_{i0}(t) + \beta\epsilon_0(t)$ is used as new input to the encoder. This results in a second set of bits. The process is repeated until a satisfactory set of bits is obtained. It should be noted that this method does not converge uniformly.

Practical results

Several laboratory models operating at speeds from 2400 bps up to 9600 bps have been built. These models have permitted a test of the key techniques described in this paper for the PCM case. Measured spectra in the case of PCM encoding are shown in Fig. 10. The data indicate that bandwidth efficiency of microcoded modem transmitters are at least as good as that of conventionally implemented digital echo modem transmitters.

Conclusion

Various microcoded modem transmitters designs have been discussed in this paper. These designs, which are based upon the use of the digital echo modulation technique with pulse-code modulation or delta-encoded signal elements, permit conventional transmitter modulators and filters to be replaced by memories and general-purpose logic circuits.

The proposed designs have been shown to be universal in the sense that most practical modem transmitters can be implemented simply by loading the proper microcode into the transmitter stores without any hardware changes.

The approach described here indicates the feasibility of the design of modem transmitters that have performances equal to or better than those of conventional modems as far as bandwidth efficiency and residual distortion are concerned. The main design limitation is the rational relationship that exists between the carrier frequency and the signaling frequency. This limitation has been shown to be mostly overcome by proper choice of the transmitter architecture and memory size.

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