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A Vestigial-Sideband, Phase-Reversal Data Transmission System

Abstract: A new method of carrier retrieval is described for a suppressed-carrier, vestigial-sideband data transmission system. Tests over voice grade telephone lines indicate that operation at 3000 bits per second (in some cases, 3600 bps) can be obtained reliably over private lines with simple adjustable equalization, whereas fixed compromise equalization will allow speeds up to 2400 bps over a large percentage of lines in the switched network. The possibility of higher speed operation using a multilevel data signal is demonstrated by tests at 6000 and 8000 bps over a carefully equalized private line.

Introduction

Double-sideband, suppressed-carrier modulation, or phase-shift keying, employing synchronous detection at the receiver has been widely accepted as a nearly optimum system for binary data transmission in terms of performance in the presence of noise.[1–4] The maximum practical bit rate of such a system is limited to about 2400 bits per second over high-quality, voice grade channels and to about 1200 bps in switched network applications, where poorer lines may be obtained and adjustable equalization is not employed.[3]

A more efficient utilization of bandwidth can be obtained either by the use of two double sideband (DSB) channels with quadrature carriers (or four-phase modulation), or by vestigial sideband transmission (VSB). Either method suffers only a modest decrease in noise performance while maintaining the inherent capability of DSB to work over a wide dynamic range of signal levels at the demodulator. A basic difficulty in both techniques has been the problem of carrier recovery at the receiver for use with synchronous detection. For certain data patterns the received signal after frequency translation in the channel does not contain the carrier information necessary to allow recovery by the methods that have been most commonly used for DSB.[2] This problem can be avoided in the four-phase system by the use of phase comparison detection.

One straightforward solution to the carrier retrieval problem has been to remove the low frequency components of the data signal by filtering, allowing a low level carrier signal to be transmitted without interference. The resulting system can be used to transmit data by employing dc restoration,[5,6] by using special signal formation,[7] or by using coding which does not require low frequency components (e.g., four-out-of-eight coding). Each of these methods has certain disadvantages:

(1) The dc restoration technique requires an accurate automatic gain control (AGC). Furthermore, the elimination of the dc component in the data signal causes a 6 dB increase in the peak line signal that occurs when the data signal changes state after being in a given state for a prolonged interval; if the peak signal determines the transmission power level, the average transmitted power must be reduced, and thus a more noise-sensitive system will result.

(2) The signal formation method which converts the original data signal into a positive pulse for each positive transition and a negative pulse for each negative transition also requires an accurate AGC. Since the transmitted signal has three levels, it is somewhat more noise-sensitive than a binary signal.

(3) The coding approach would be practical only in an over-all data system where the serial data are readily provided in the desired form.

This paper presents a new solution to the carrier re-

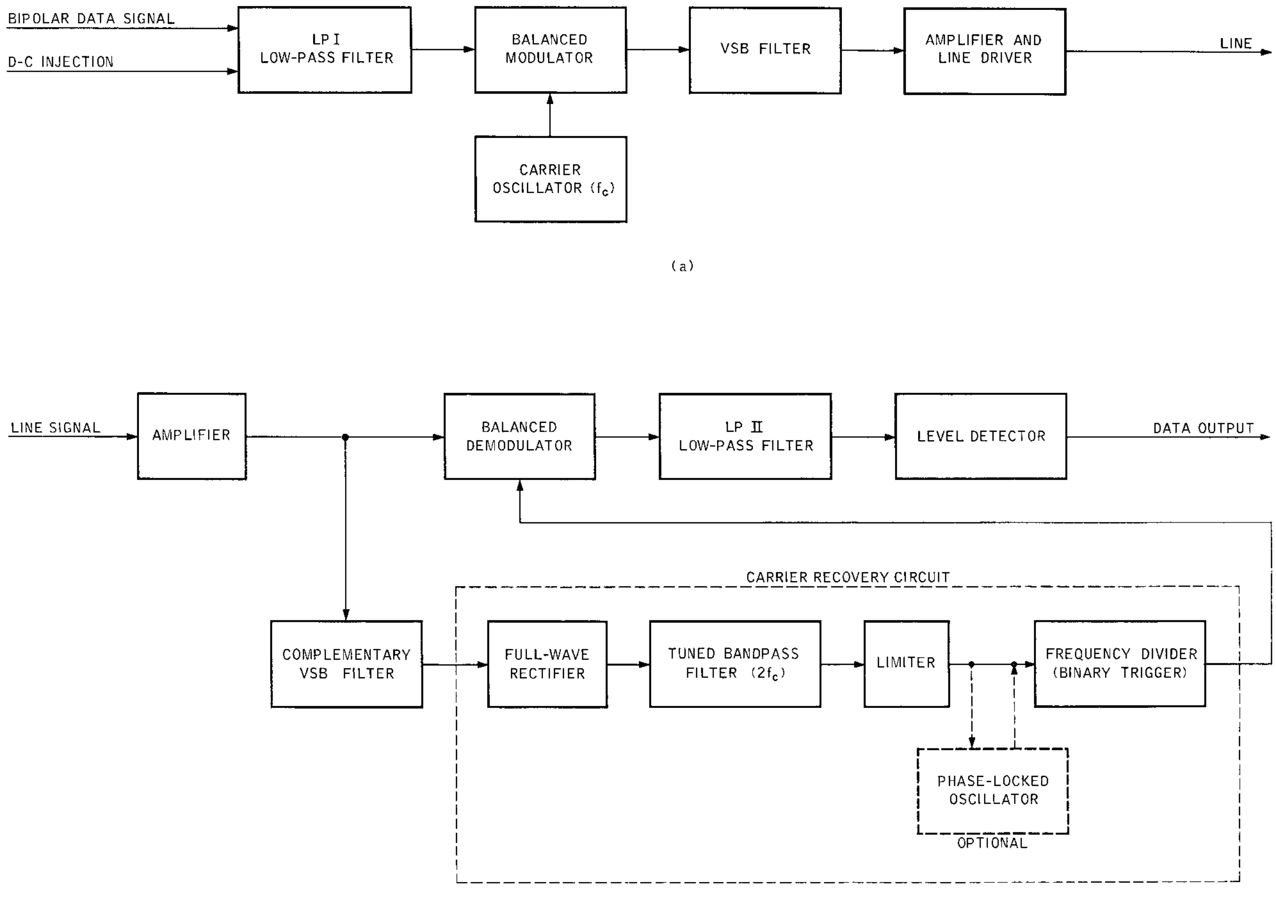


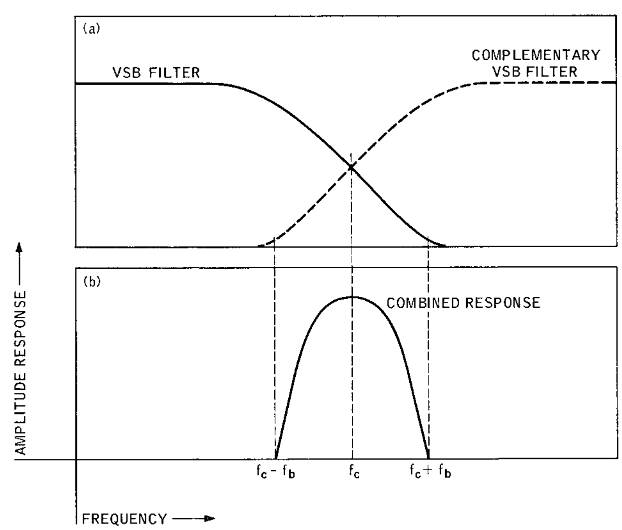
Figure 1 **Vestigial-sideband data transmission modem.** (*a*) *Transmitter;* (*b*) *receiver.*

covery problem, leading to the development of a VSB modulation-demodulation system (modem) which retains the inherent simplicity and reliability of DSB phase reversal modems. This modem allows synchronous transmission (continuously clocked) or asynchronous transmission (clocked during each character, e.g., teletype data) with no coding restrictions on the data. No synchronization between the data and the carrier is required. The system may be used to transmit binary data directly or with multilevel encoding to increase the information rate.

Carrier recovery for a vestigial sideband system

The VSB system shown in Fig. 1 is identical to a DSB phase reversal system to which has been added the VSB and complementary VSB filters. These filter responses are shown in Fig. 2a and have the combined response of Fig. 2b. The transmitted signal spectra for DSB and VSB are illustrated in Fig. 3. In the case of DSB the total signal

Figure 2 **Characteristics of the VSB filter and of the complementary VSB filter.** (*a*) *Their individual responses;* (*b*) *their combined response.*



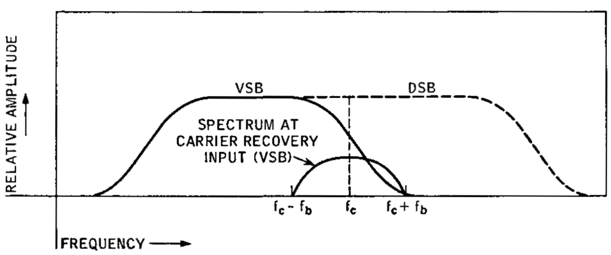
spectrum is applied to the carrier recovery circuit (shown in the dashed lines of Fig. 1b) which essentially beats the upper and lower sidebands together to generate an amplitude modulated tone at $2f_c$, which does not reverse in phase. The Appendix presents a comprehensive analysis of this process.

In the VSB system only the partial signal spectrum of Fig. 3 (consisting of the DSB spectrum modified by the combined filter response of Fig. 2b) is applied to the carrier recovery circuit. Thus the carrier is retrieved from signals in this spectrum which correspond to components of the baseband data signal in the frequency range 0 to f_b. The difficulty which has prevented prior utilization of this system is that certain data signals (e.g., a repetitive 1–0–1–0 NRZ bipolar signal) contain neither dc nor components lower in frequency than f_b. It will be shown next that this difficulty can be overcome by the addition of a small dc component of carefully chosen magnitude to the bipolar data signal. (This is equivalent, of course, to adding a small carrier tone to the suppressed carrier output of the balanced modulator.)

In a practical case the number of repetitive patterns containing no frequency components less than f_b is fairly limited. A pattern repetitive in N bits has a lowest (fundamental) component at frequency f_d/N, where f_d is the data rate. Consider as an example a case in which $f_b/f_d = 1/5$ (corresponding to a VSB filter passing up to $f_c + 600$ for a data rate of 3000 bps). Patterns of $N = 6$ and greater have fundamental components which allow carrier recovery. The possible dc averages of all possible 1, 2, 3, 4, and 5 bit patterns may be seen by inspection to be 0, $\pm\frac{1}{5}$, $\pm\frac{1}{3}$, $\pm\frac{1}{2}$, $\pm\frac{3}{5}$, and ±1 v, assuming the data signal swings between ±1 v. If a dc level of 1/10 v is added to the data signal, the smallest average for any pattern of 5 bits and less becomes $\pm1/10$ v. It should be noted that the value of the injected dc, being half the lowest nonzero average value for all the patterns, is an optimum choice since any other value (except for values greater than 0.7) would yield a smaller net dc for at least one data pattern. This rule should also be followed when designing for a different value of f_b/f_d. (For $f_b/f_d = 1/3$, the dc injection should be 1/6, for example.)

It has been shown that all possible data patterns can be assured of having components for carrier recovery at the receiver. The magnitude of these components, as shown by experiments, is sufficient that the recovered carrier signal is at least as strong (for any data signal) as when the worst-case 1–0–1–0 $\cdots$ pattern is transmitted. Thus the signal level at the input of the carrier recovery circuit varies over a range of 11 to 1 for a random data signal with the dc injection value of 1/10.

Figure 3 **Line signal spectrum for VSB and DSB.**



The low level of carrier for some data patterns suggests that the recovered carrier might be overly sensitive to channel noise. However, the carrier recovery system is responsive to noise only in a portion of the bandwidth. When the tuned circuit of Fig. 1b has a Q of 30, the degradation of system performance for a random data pattern is about 1.5 dB because of noise on the carrier compared to performance with a perfect patched-through carrier (as found in the random noise tests described in the next section). It is possible to reduce the degradation to about 0.5 dB by adding a phase locked oscillator, as shown in Fig. 1b.

Other methods are possible for operating on the data signal to insure the presence of low frequency components. These will be described briefly:

(1) A variable bias distortion can be introduced in the data signal, e.g., by delaying positive-going transitions. This causes an additional dc component that is proportional to the rate of transitions. Thus the patterns with no low frequency components, which must contain many transitions, obtain the largest dc injection.

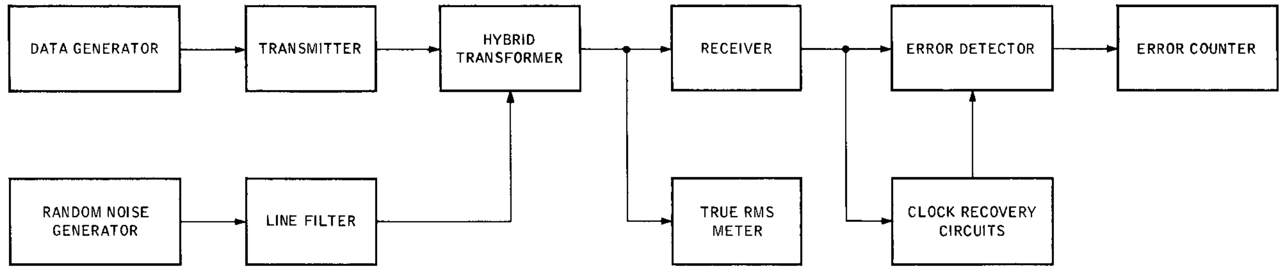
(2) The baseband data signal can be amplitude modulated (at a low modulation level) by a repetitive pattern such as 1–0–0–1–0–0 $\cdots$. Low frequency modulation products are produced for those patterns which did not have low frequencies previously.

Both of these methods have been investigated analytically and experimentally and found to give performance similar to that for constant dc injection. Because of its simplicity, constant dc injection appears to be the most practical technique.

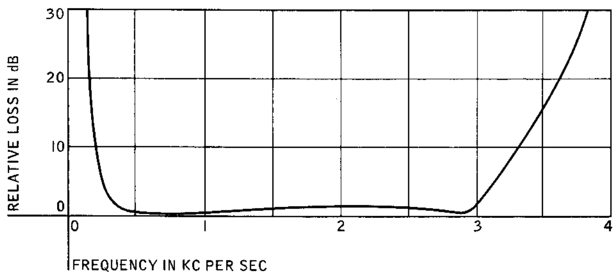
Binary transmission

Several types of binary transmission systems can be designed using the basic modem of Fig. 1. One of the several common methods of deriving a data clock from the data signal may be used at the receiver. Although the analysis above concerns a bit-synchronous system (i.e., using a continuous clock), further analysis shows that the carrier retrieval will also operate successfully for start-stop transmission, as, for example, using the standard teletype code.

One characteristic of the modem must be given consideration: The data output at the receiver may be inverted if the bistable recovered carrier is started in the wrong phase or if it reverses phase during a line disturbance. This may be resolved by coding the information



(a)



(b)

Figure 4 **Experimental arrangement for binary noise tests.** *(a) Test setup to obtain relation between error rate and signal/noise ratio; (b) attenuation characteristic of line filter.*

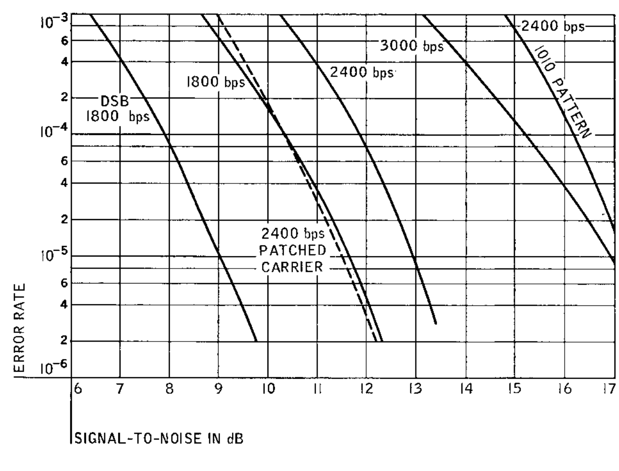
in level transitions rather than in absolute levels. An alternative is to provide circuits to synchronize the recovered carrier at the beginning of each message. Simple circuits have been devised which perform this function reliably.

- *Noise tests*

Tests performed in the presence of random noise do not give an absolute measure of the reliability to be expected when operating on actual telephone lines. Nevertheless, such tests are indicative of reliability relative to that of other systems tested under similar conditions. Figure 4 shows a test setup which was used to obtain the relationships between error rate and signal-to-noise ratio (S/N) shown in Fig. 5. The noise from a flat gaussian source is band limited by the simulated line characteristic of Fig. 4b, which is of sufficient bandwidth to pass the transmitted signal spectrum; however, the transmitted signal has been connected directly to the receiver to avoid bringing a small phase impairment into the experiment. The carrier recovery uses a tuned circuit with a Q of 30 for the $2f_c$ filter (with no phase-locked oscillator), and the dc injection is 1/10 of the data signal level as used in the example of the previous section. The data clock is reconstructed at the receiver by circuits that effectively average the transition times of the output data waveform.

Comparison of the results at 2400 bps with those for a perfect (patched through) demodulation carrier shows a 1.5 dB difference due to the noise-produced phase jitter on the recovered carrier, for the case of a quasi-random data pattern. The result for a 1–0–1–0 $\cdots$ repetitive data

Figure 5 **Error rates for binary transmission with band-limited gaussian noise.**



signal shows an additional 4 dB decrement when this "worst case" pattern is transmitted continuously.

Results at 1800 bps show that the VSB system is 2 or 3 dB more noise-sensitive than a DSB system at this same speed. The DSB system used the same components except for removing the VSB filters and retuning the carrier.

Theoretically the results at 3000 bps should be only about 1 dB different from those at 2400 bps. Jitter in the recovered data clock caused a 1-dB decrement which was not present at the lower speeds. A slight amount of intersymbol interference, caused by band-limiting in the data filters, was also present at 3000 bps.

• *Line performance*

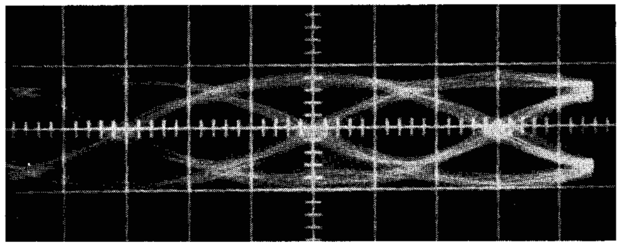
Tests over a variety of actual leased lines and simulated lines have shown that in most cases 3000 bps transmission can be achieved using quite simple equalization techniques. A carrier frequency of 2500 cps has been found to give good results with a transmitted spectrum of about 500 to 3000 cps. Figure 6a illustrates the results that were achieved over a long-distance line, showing the "eye diagram" at the output of filter LP-II of the modem (a superposition of the system response to a quasi-random data sequence). This line, consisting of 3 carrier links and 20 miles of loaded cable, was equalized by two adjustable, second-order equalizer sections of conventional design.

In many cases line quality is sufficient to allow greater transmission speed using a higher carrier frequency. Operation at 4000 bps over a single equalized high-quality carrier link is shown in Fig. 6b.

For dial-up operations where adjustable equalization may not be economically feasible, a fixed compromise equalization may be used. Maximum practical speeds can be determined only after exhaustive tests over a large number of lines and will vary from country to country. Tests so far indicate that speeds in the order of 2000 to 2400 bps can be expected except over lines with long lengths of heavily loaded cable. As an example, the modem has been tested over a simulated line of three carrier links and 50 miles of H88 loaded cable. Figure 7 shows the characteristics of this line along with those for a compromise equalizer. Figure 8a shows the eye diagram at 2400 bps (with a carrier of 2300 cps) over this line using the compromise equalizer. The back-to-back response through the compromise equalizer, shown in Fig. 8b, is also adequate. The oscilloscope was synchronized by a data clock, derived from the received data (also shown in Fig. 8).

Quaternary transmission

In the switched telephone network, where one has little or no control over the quality of the line, it is necessary to choose a data modem with wide margins of protection against noise and distortions. On the other hand, private line systems are normally of higher quality since it is



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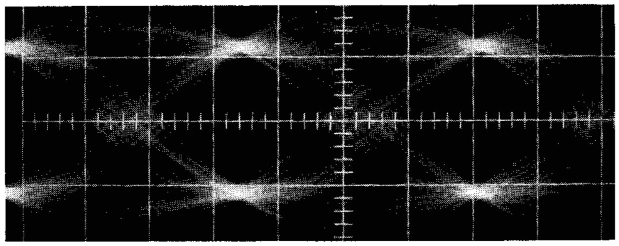
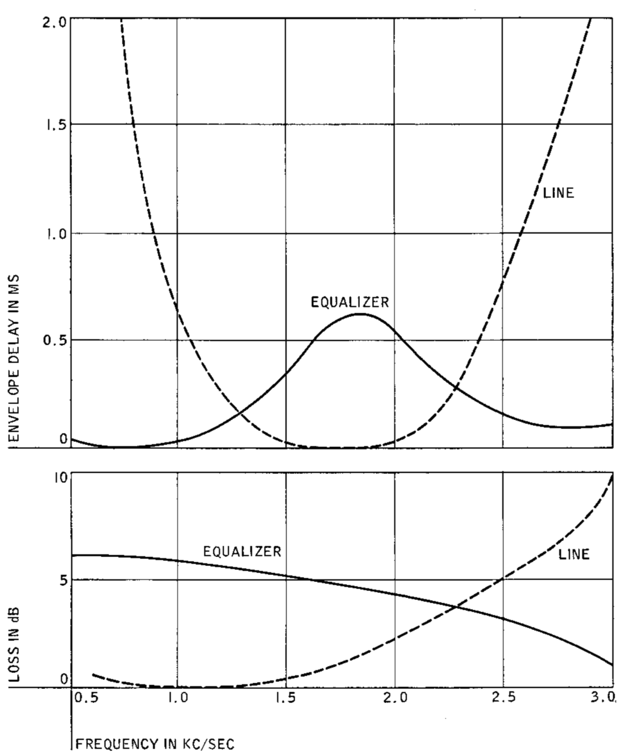
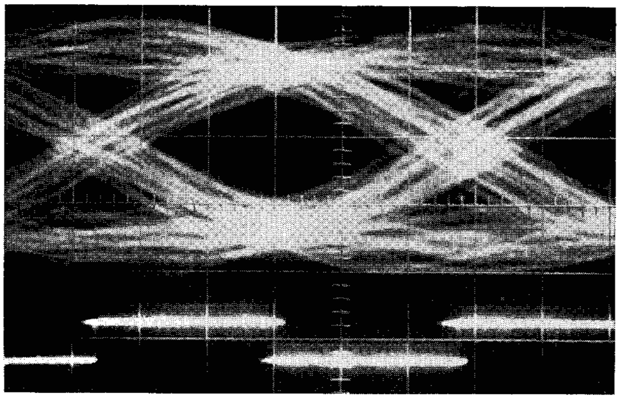


(*b*)

Figure 6 **"Eye diagrams" for binary operation.** (*a*) *3000-bps operation on long-distance line;* (*b*) *4000-bps operation on single carrier link.*

Figure 7 **Amplitude and phase characteristics of compromise equalizer and simulated line.**





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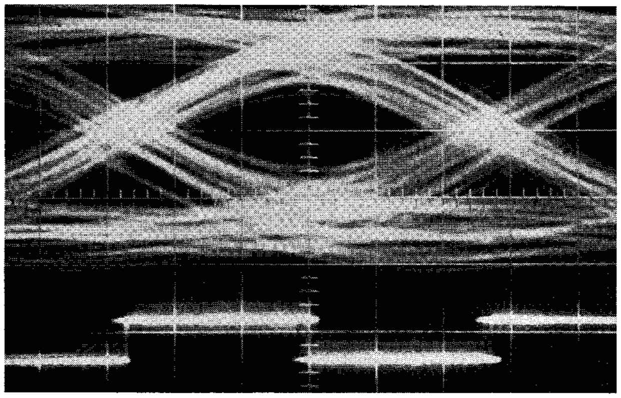


(*b*)

Figure 8 **"Eye diagrams" using fixed compromise equalization (synchronized to receiver clock).** *(a) For operation over simulated line of Fig. 7; (b) for back-to-back operation.*

possible to study the route, equalize the circuit and control the noise to a certain extent. A signal-to-noise ratio of at least 30 dB can be expected, and useful bandwidths may be as large as 2800 to 3300 cps.

It is possible to take advantage of improved line conditions to provide higher transmission rates by encoding the binary data into a multilevel waveform that can be transmitted by the VSB modem, since it is essentially a linear system from the input of the data filter (LP-I) at the transmitter to the output of the filter LP-II at the receiver. The only change required in the basic VSB modem is to set the dc injection level to account for the new average dc levels of the multilevel waveform. Although the use of multilevel coding is not novel,[8] it deserves further investigation to enable us to explore the speed limitations of presently available transmission systems using the VSB modulation techniques described here.

Four-level transmission has been chosen for investigation since it appears to yield a worthwhile increase in speed without becoming unduly complex. Information on two synchronous binary signal lines is applied to the four-level encoder which produces an output in accordance with the table of Fig. 9. This coding takes on physical meaning when it is observed that Signal A controls the polarity of the four-level signal while Signal B controls the magnitude (corresponding to the phase and amplitude of the modulated signal). The optimum dc injection at the transmitter (from calculations similar to those previously shown) is 1/3 the value which would be optimum for a binary signal at the same digit rate, i.e. at half the bit rate of the four-level signal. At the receiver the four-level waveform at the output of LP-II is sampled at the center of each digit to determine the amplitude and polarity for reconstruction of the two binary signals.

- *Functional tests*

Noise tests for four-level transmission at 6000 and 8000 bps have been performed using the setup of Fig. 10. In this case the transmitted signal and the injected white noise were both passed through a carefully equalized carrier channel. The clock source at the transmitter was also used at the receiver to determine sampling time. Two sets of results are shown in Table I, where one set is for a patched-through carrier and one is for the same carrier retrieval circuits used for the binary tests. The noise decrement resulting from the carrier retrieval method (the difference between these two sets of results) indicates the need of a phase-locked oscillator for operation at these high rates. The performance of the phase channel of the four-level system compared to binary operation at the same bit rate is seen to be only 6 dB worse on an average power basis.

The eye diagram of the received four-level waveform for the 8000 bps test is shown in Fig. 11b. The carrier frequency was 3000 cps. (Binary operation over this same line was shown in Fig. 6b.) Figure 11a shows four-level operation at 4000 bps over the long-distance line which was previously described for the binary test of Fig. 6a. An adjustable tapped delay-line equalizer was used to obtain optimum waveforms for both these cases.

- *AGC and clocking operations*

The practical implementation of a four-level system requires the addition of a carefully designed AGC system

Figure 9 **Function of the four-level encoder.**

A	B	OUTPUT
1	1	+V
1	0	+V/3
0	0	-V/3
0	1	-V

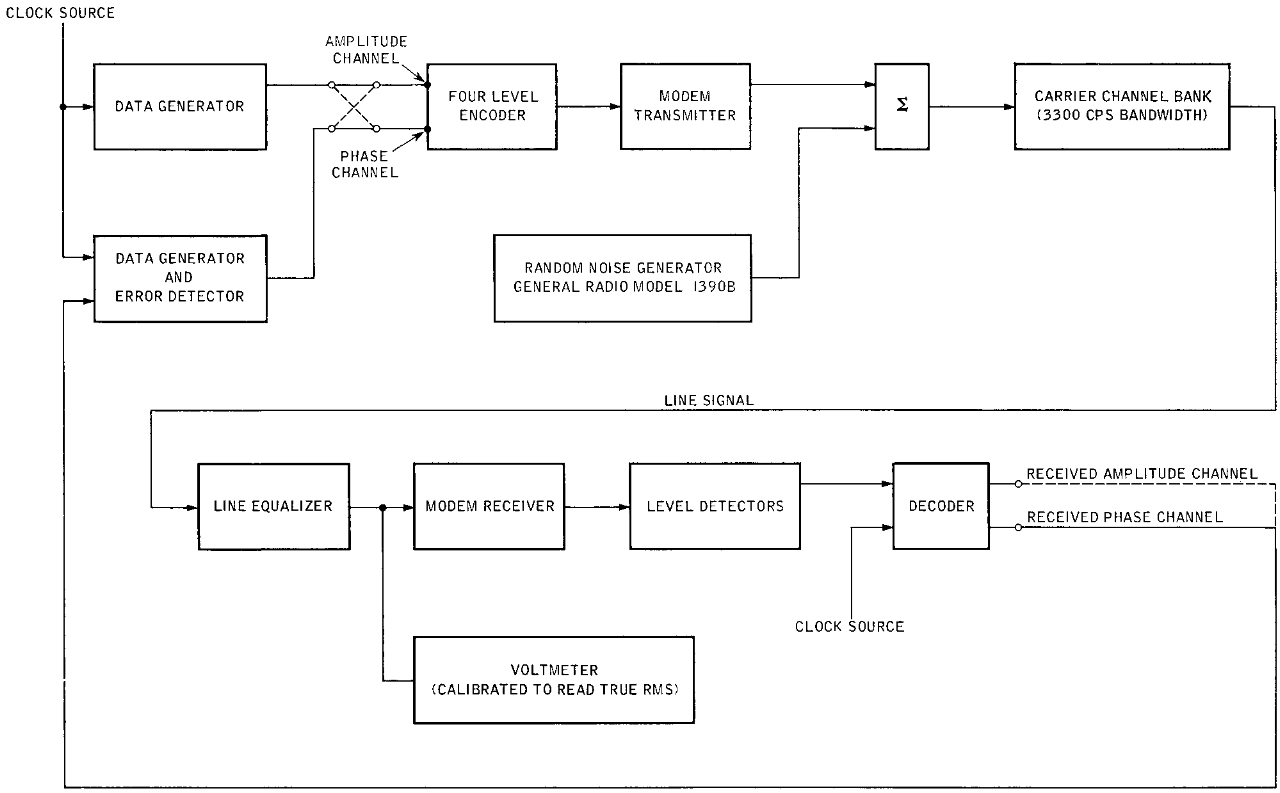
and places more stringent requirements on receiver clocking. The AGC must be accurate to about ± 1 dB and cannot be regulated from the power level of the line signal since it is not constant. The clock problem arises because the time at which the data waveform crosses a given threshold depends on the level of the previous digit and the level of the next digit. As a result the horizontal opening of the eye diagram is generally less than for binary operation and the sampling time must be more carefully controlled. The recovery of the clock from the data wave-

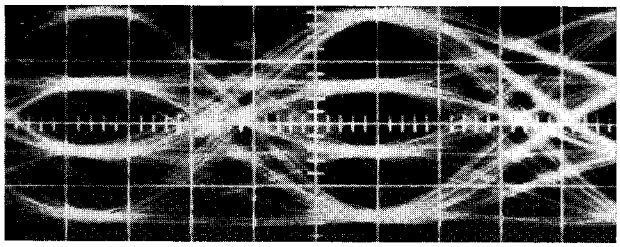
Table 1 **Modem noise test results (clock patched in all cases).**

	S/N Ratio in dB for 10^{-5} Error Rate					
	Four-Level*				Binary	
	6000 bps Phase Channel	6000 bps Amplitude Channel	8000 bps Phase Channel	8000 bps Amplitude Channel	3000 bps	4000 bps
Carrier patched	19	20	20	24	13	14
Carrier recovered	22	23	27	30		

* Phase and amplitude channels of a four-level system measured separately.

Figure 10 **Experimental arrangement for quaternary noise tests.**





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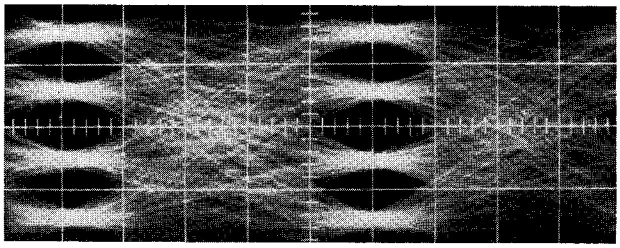


(*b*)

Figure 11 **"Eye diagrams" for quaternary operation.** *(a) For 4000 bps on the long-distance line; (b) for 8000 bps on a single carrier link.*

form is also more difficult because of this inherent threshold-crossing jitter. However, the time at which the waveform experiences maximum slope during a level transition is independent of the type of transition (except for interbit distortion). It is possible to detect this maximum slope by a differentiation process, giving a signal that may be used to synchronize a clock in the receiver. This scheme may be used for clock recovery when sufficient transitions are present in the data waveform.

One solution to both the AGC and clocking problems involves the transmission of a pilot frequency outside the normal line signal spectrum at a frequency $f_p = f_c + f_s/n$, where f_s is the sampling (digit) rate. The clocking information is derived at the receiver from the pilot tone and the recovered carrier by multiplying their difference frequency by n to obtain $f_s = n(f_p - f_c)$. (Note that spectrum shifts of f_p and f_c cancel out in this process.) As an example, for 4800 bps operation a carrier of 2400 cps and $n = 4$ could be chosen, resulting in a pilot tone at 3000 cps. The power level of this same tone can be monitored at the receiver and used to regulate an AGC. In practice it has been found convenient to regulate the detection thresholds automatically to achieve the same effect.

Several other solutions to these problems have been postulated but will not be presented here. Further investigation is necessary to determine their relative merits.

Conclusions

The new vestigial-sideband data transmission system which has been described in this paper has certain desirable characteristics which are found only partially in previously described systems. Data can be transmitted with no coding restrictions and without synchronization to the carrier; asynchronous data such as teletype can also be handled. Binary transmission does not require the use of an AGC for private line applications where the gain variation is less than ± 4 dB, and only a simple AGC is required for switched-network applications. The ratio of peak transmitted power to average power is only slightly greater than unity.

Furthermore, the basic system can be extended for quaternary transmission over high quality lines to increase the data rate. The quaternary system is somewhat more sophisticated, requiring careful amplitude and phase equalization and a precise AGC.

Although detailed circuit diagrams for implementing the described system are not presented here, it should be noted that several models have been built. The reliability of all elements of the system has been verified in more than a year of laboratory and field experiments.

Acknowledgments

The authors wish to acknowledge the valuable contributions of G. K. McAuliffe, formerly of IBM, in his initial efforts towards proving the feasibility of vestigial-sideband, phase-modulation systems. J. S. Chomicki has built and tested almost all the circuits, and thus has played a very important part in the experimental work and has substantially contributed towards the success of the entire effort.

Most of this work has been carried out at the East Coast Laboratory of the IBM Advanced Systems Development Division in Yorktown Heights, New York. Part of the experimental work was performed at the Centre d'Études et de Recherches, IBM France, La Gaude.

Appendix: Carrier recovery from sideband components

The only way to recover the carrier in the correct frequency and phase relative to the sidebands after going through a system having spectrum shift is to send the carrier itself, or to send two or more tones which are related to the carrier in some special way, so that they can be used to derive the carrier. A simple way to generate such a pair of tones is to modulate the carrier with a tone at baseband, resulting in an upper and a lower sideband component, symmetrically located about the carrier. In fact, when the carrier is modulated with the baseband data signal at the transmitter, the result is a number of pairs of components spaced about the carrier. If the signal is sent on the line as DSB, the condition of at least two components symmetrically spaced about the carrier is always satisfied, assuming that the data signal does not rest at zero for any appreciable length of time (e.g., it may be bipolar NRZ). A method of using these sideband components to retrieve the carrier is illustrated in Fig. 12. This method can be partially understood by considering the carrier wave $\cos \omega_c t$ to be modulated by a signal $\cos(\omega_0 t + \theta_0)$. When this modulation product is transmitted in a channel with amplitude response $A(\omega)$, phase response $\phi(\omega)$, and a small frequency translation α, the received signal is given by

$$V(t) = \frac{A(\omega_c + \omega_0)}{2} \cos[(\omega_c + \omega_0 + \alpha)t + \theta_0 + \phi(\omega_c + \omega_0)] + \frac{A(\omega_0 - \omega_0)}{2} \cos[(\omega_c - \omega_0 + \alpha)t - \theta_0 + \phi(\omega_c - \omega_0)]. \tag{A-1}$$

The output of the square law device of Fig. 12 is

$$V^2(t) = \frac{1}{4}\left\{ \frac{A^2(\omega_c + \omega_0)}{2} \cos 2[(\omega_c + \omega_0 + \alpha)t + \theta_0 + \phi(\omega_c + \omega_0)] + \frac{A^2(\omega_c - \omega_0)}{2} \cos 2[(\omega_c - \omega_0 + \alpha)t - \theta_0 + \phi(\omega_c - \omega_0)] + A(\omega_c - \omega_0)A(\omega_c + \omega_0) \cos[2(\omega_c + \alpha)t + \phi(\omega_c + \omega_0) + \phi(\omega_c - \omega_0)] + \text{terms at dc and } 2\omega_0 \right\}. \tag{A-2}$$

The dc component and the $2\omega_0$ component are easily filtered out. The remaining three components are shown by phasors in Fig. 13 for the conditions

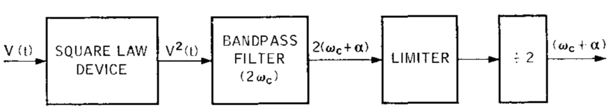
$$A(\omega_c + \omega_0) = A(\omega_c - \omega_0),$$

and $\phi(\omega_c + \omega_0) + \phi(\omega_c - \omega_0) = 2\phi(\omega_c)$,

i.e., the amplitude is even symmetric and the phase is odd symmetric about the carrier. In this case the zero crossings of the resultant waveform, Fig. 14, are not affected by the sideband components. If the conditions of symmetry above were not met, there would be a phase modulation of the waveform causing a jitter in the zero crossings of the $2\omega_c$ carrier.

This waveform, Fig. 14, can now be passed through

Figure 12 **System for carrier retrieval from sideband components.**



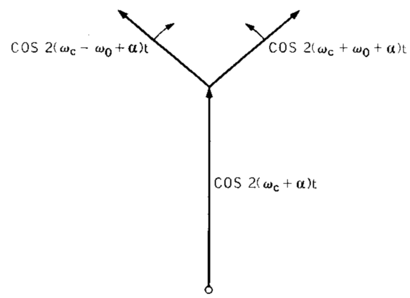
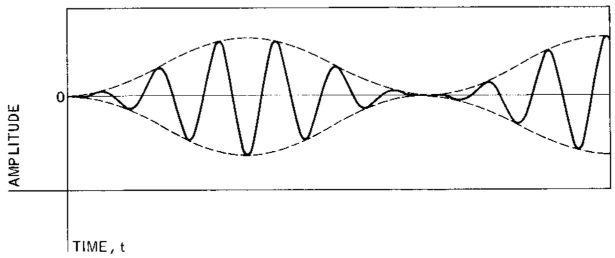


Figure 13 **Phasor diagram of square-law device output.**

Figure 14 **Waveform produced by phasor resultant.**



a narrow bandpass filter which is symmetrical about $2\omega_c$. The degree to which this filtering reduces the modulation at $2\omega_c$ is dependent upon the bandwidth of the filter compared to the modulating frequency. (The phase characteristic of the bandpass filter must be such that a change of the spectrum shift does not cause an appreciable phase change in the recovered carrier. If a tuned circuit is used, the maximum Q is determined by the maximum spectrum shift of the channel, which is normally less than ± 15 cps.) The remaining amplitude variations are removed by a limiter that clips the waveform at the zero crossings. The limiter output is divided down to obtain the desired carrier frequency for synchronous demodulation. (A constant phase shift may also be necessary to compensate for a phase shift in the division process.) An ambiguity in polarity of the carrier is introduced by the divider.

It would appear that as the modulation frequency is reduced more and more the Q of the tuned circuit would have to be increased. However, in a data system the modulating waveform contains harmonics which prevent the modulation envelope of $2\omega_c$ from staying near to zero for long periods of time. This is most easily analyzed by another approach. Note that if the channel is distortionless, the output of the square law device can be written as

$$V^2(t) = [V_0(t)]^2[\cos(\omega_c + \alpha)t]^2, \tag{A-3}$$

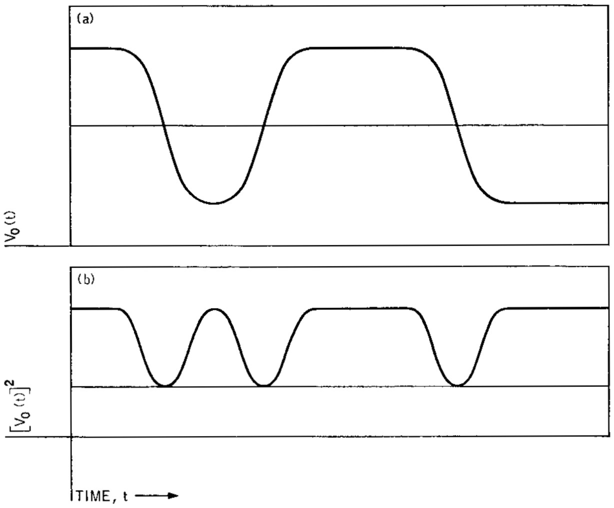


Figure 15 **Filtered data waveform and the square of this waveform.**

where $V_0(t)$ is the filtered data signal at the input of the modulator.

Expanding the cosine term gives

$$V^2(t) = \tfrac{1}{2} V_0(t)^2 \cos 2(\omega_c + \alpha)t + \tfrac{1}{2} V_0(t)^2. \qquad \text{(A-4)}$$

The essential input to the carrier bandpass filter is, thus, a carrier of frequency $2(\omega_c + \alpha)$ modulated by the square of the original modulating signal. The filtered data waveform and its square for a bipolar NRZ signal are illustrated in Fig. 15 (*a* and *b*). Note that the square of the waveform consists of a large dc term plus a small amount of harmonics which are present because of the premodulation filtering. This waveform is the envelope of the $2\omega_c$ carrier at the input to the bandpass filter. Thus the filtering problem is not great.

A similar analysis is applicable to the VSB system. Figure 16 illustrates envelopes of the waveform at successive stages of the VSB carrier retrieval system. The modulator output shown in Fig. 16a has an envelope which is the filtered data signal including the injected dc component. At the input to the carrier retrieval circuits this signal is reduced in bandwidth, as shown in Fig. 16b, by the low-pass equivalent of the combined VSB and complementary VSB filter responses. (Symmetry of the combined response of Fig. 2b prevents phase modulation of the carrier at this point, assuming that the channel produces no distortion.) The output of the full-wave rectifier, which in practice replaces the square law device, is shown in Fig. 16c. The tuned circuit output at frequency $2f_c$ has the envelope shown in Fig. 16d. It has also been found desirable in practice to clip the output of the full-wave rectifier to reduce the amplitude range of the signal applied to the tuned circuit.

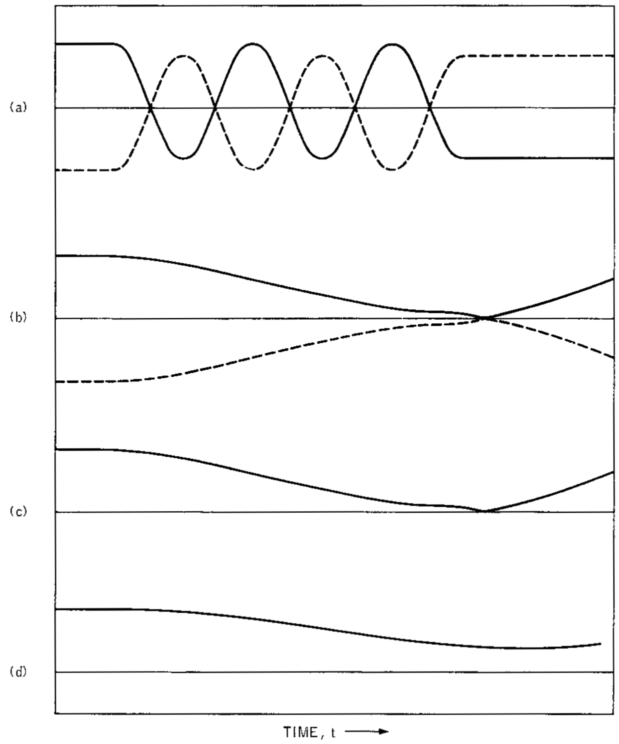


Figure 16 **Envelopes of waveforms illustrating VSB carrier retrieval.**

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