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TITLE: **RF MODULATION PROPOSAL: QUADRATURE DOUBLE SIDEBAND REDUCED CARRIER WITH TWO NRZST BASEBAND CHANNELS**

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SUMMARY

A two part modulation is offered for consideration which consists of: 1) up-down conversion between baseband and QAM modulated radio frequency, and 2) I and Q channel smoothed (or sinusoidal transition) NRZ baseband modulation. The resulting radio signal resembles QPSK (two superimposed PSK) retaining polar detectability and but without out-of-band components that are generated by non-linear methods.

This modulation is thought suitable for spread spectrum at a 30-40 Mb/s chipping rate or for other code forms at higher rates. The required radio is linear with a spectral density of 2 bits/Hz. In the processing details, this form may have significant advantages for this application.

For reduced cost and power-drain, only one of the two available quadrature phased channels may be used resulting in simpler PSK.

Attachment A is an edited and condensed version of the January 1990 paper presenting NRZST to IEEE 802.9 for use at 16 Mbs (20 Mb/s with 4/5 line coding).

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GENERAL

There has always been an argument between advocates of constant envelope and linear modulation for digital transmission. Constant envelope has generally prevailed in land mobile and some satellite, but linear has become predominant in point-to-point microwave and applications demanding minimum use of spectrum.

Many of the arguments have been in the context of analog voice transmission, and for narrowband systems using premises that are not applicable to wireless LAN. It is the intention to offer the case for an obvious linear modulation, so that it may be evaluated against well understood constant envelope methods.

The proposed modulation is not constant envelope, and it benefits from linearity in the radio amplifiers. It is described from the **viewpoint of independent I and Q components rather than resulting composite signal.**

For reduced cost and power-drain, only one of the two available quadrature phased channels may be used resulting in simpler PSK.

Context

The assumed context is 30 to 40 Mb/s which might be the chipping rate for a spread spectrum system of 1-1.5 Mb/s data throughput or other possibilities. Simplicity and low power drain are also sought.

Assumptions

One of the more subtle assumptions is that FM is insufficiently spectrum efficient for data rates of 10 Mb/s up when the modulation index is high enough to give sufficient quieting advantage.

Another assumption is that the difference in efficiency between limiting and linear transmitter power amplifies is unimportant when only 1.0-2.5 milliwatts of transmitter power is required.

Another assumption is that the two-tone test for linearity does not need to have the same degree of suppression of intermodulation products that applies to higher power transmitters.

Propagation

Propagation studies and analysis by Masleid, Van der Jagt and Storoshchuk show phase continuity of the received signal for 100's of bits at high megabit rates. Fading is slow compared with the symbol period for a high bit rate.

DESCRIPTION

The modulation is seen as being in two parts: 1) the up-down and down-up converter between two quadrature baseband analog signals and the radio frequency, and 2) the digital signal processing of these two baseband signals for receiving and transmitting.

Reference block diagrams for the up/down converter are shown in contribution 802.11/91-99.

Baseband Waveform

The baseband signal is NRZ with sinusoidal transitions (NRZST) the spectrum width of which is half the bit-rate above dc. This modulation was studied and models built for IEEE 802.9 transmission on unshielded twisted pair at 16 Mb/s. It was found that a spectrum could be produced with negligible energy above the first zero by generation with an FIR filter with a window of five symbols. A rectangular or sinusoidal waveshape at baseband would have a continuous series of lobes at lower amplitudes beyond the necessary bandwidth for information transmission.

A main objective in the design of the transmitting circuits is retaining the narrow spectrum of the baseband signals. This requires good linearity of the multiplier and amplifiers in the transmitter.

Up Conversion

The transmit signal is formed with two four-quadrant linear multipliers. One input for both is the rf carrier frequency with 90° phase difference and the other is two independent data streams with NRZST shaping. For QPSK, the binary input amplitudes are +1 and −1. For QAM, they would be +1 and 0. Only one of the two multipliers would be used for PSK.

Offsetting to obtain near constant envelope is not introduced. The signal will have an envelope similar to DSBSC. In part, this is possible because limiting will not destroy the information content at the receiver, since the information is carried in the phase and not the amplitude of the signal. The envelope is an aid in fast bit clock recovery.

The purpose of the NRZST is to define each data pulse over several pulse intervals to control the out-of-band spectrum generated and to reduce inter-symbol effects to negligible levels.

The resulting signal is very similar to PSK except that by looking at it as a QAM linear transmission path with low spectrum baseband signals, a narrow spectrum rf signal is obtained without using band filtering.

Carrier Insertion

It is possible to transmit full or partial carrier so that rf reference carrier may be more quickly recovered than if it is synthesized. There are probably several different implementation methods that would provide a sufficient result.

The carrier should be inserted with a separate calibratable injection circuit.

It is possible to create a carrier component by introducing dc into the baseband signal on one of the two phases applied to the multipliers. This method is more reproducible than unbalance. The carrier injection can be maintained for just the preamble or for the entire duration of the transmission.

Down Conversion

Before the inverse process can take place, it is necessary to locally generate a replica or approximation of the carrier frequency used to generate the signal. One obvious method is to transmit reference carrier and recover it, but there are many known alternatives. Some possibilities are:

1) Using inserted carrier reference in the transmission, the desired frequency can be recoverd with narrow band filtering.

2) The carrier may be synthesized using the redundancy in double sidband.

3) The reference may be the phase of the preceding symbol delayed.

4) The incoming signal may be squared and then divided.

5) If the local oscillator is close to the incoming carrier but not phase locked, the resulting precession in the baseband signal from frequency difference may be taken out with digital signal processing of the baseband signal.

32 Mb/s rate in the medium results in a detection noise bandwidth with matched filtering of 8 MHz in each channel.

If the triple send for 10+ Mb/s is linearly combined, the noise bandwidth can be reduced to less than 3 MHz in each channel. Simpler circuits may not do as well.

The inherent characteristic of linear modulation is the signal-to-noise ratio at and after detection cannot be better than the signal-to-noise ratio of the original signal. With binary information, this is much less of a burden than it would be for voice.

SPECTRUM REQUIREMENTS

If the signal is generated with two quadrature phased multipliers, the baseband signal in each branch (I and Q) is NRZ at half the total bit rate. Assuming that spread spectrum with a chipping rate of 32 Mb/s is used, then each baseband line is carrying half of the information at 16 Mb/s with most energy below 9 MHz.

This type of modulation will have the same spectrum at signal frequency (mirrored on both sides of the carrier) as at baseband. For 32 Mb/s in the medium, the 3 dB down points will be spaced 16 Mb/s. Without strain, this would fit in a 40 MHz channel allocation (but not in less than 30 MHz).

RADIO DESIGN CONSIDERATIONS

There are a number of inter-related radio design considerations that are affected by the details of the modulation choice. A number of possibilities are mentioned.

Receiver AGC Method

Usable receiver input signal levels may vary over a range of 80 dB. The allowable time for gain adjustment is at most a few bits, a speed not achievable with ordinary automatic gain control.

Since the detection is polar and based on phase rather than amplitude, there is not a threshold setting problem. The main consequence of receiver non-linearity would be crosstalk coupling between I and Q phase and between consecutive symbols. These effects would cause the most degradation at the lowest signal levels near the threshold.

It is believed that if the receiver is optimized for linear operation for the first 20 dB above threshold, that instantaneous soft limiting may be introduced to cope with higher levels. At high signal-to-noise ratios, the phase information in the signal will not be significantly degraded by residual crosstalk that is 20 dB down.

Instantaneously, the limiting will hold levels adequately at the demodulator leaving time for the AGC to work, and the time required for operation is much less for 20 dB of linear AGC range than for larger values.

Carrier Recovery Methods

If carrier is transmitted, recovery is possible with filtering. This is not as easy as it might seem because the phase delay in the filter is likely to be frequency dependent unless especially designed.

A direct carrier recovery circuit is not obvious or easy when the speed of acquisition and noise resistance needed are taken into account. The information is obtained from the redundancy in double sideband operation.

The "Costas" loop will deduce carrier frequency from the sidebands, but may take some time to reach a decision. The result is independent of data patterns.

If the composite signal is applied to a squaring circuit, the result is independent of the phase reversals. A divide-by-two must be used to recover the carrier. The two-mode ambiguity in recovered phase would make no difference in a system where the line coding is whether the current pulse is the same or reverse phase relative to the preceding pulse. Squaring applies to binary PSK. Quadrupling and divide-by-4 would be required for QPSK.

This operation would have to be applied at an intermediate frequency in a conversion chain.

Carrier may be transmitted directly during the preamble, acquiring lock and coasting thereafter. This might be very satisfactory where a small processing difference is unwound by the processing of the baseband signal.

System Frequency Matching

To have all stations on the same radio frequency as closely as possible, the reference could be any or a selected Access-point. Correction schemes with memory are possible that would diminish the error every time a reference quality signal was received.

ATTACHMENT A
NON-RETURN TO ZERO SINUSOIDAL TRANSITION MODULATION

BACKGROUND AND NOTES

This Attachment has been derived by editing of Document No. IEEE 802.9-90/7 offered on January 11, 1990. The conceptual work on implementation with FIR pulse generation is that of G. L. Somer.

- On a telephone twisted pair medium, the NRZST technology provides both 1.6 times the data rate and 2.0 times the distance, at least, compared with 10BaseT. The susceptibility to impulse noise is probably similar to or better than 10BaseT and better than any multi-level system at the same data rate.

- One of the important insights of this signaling method is that it is better and easier to compensate for predictable phase and amplitude distortion at the transmitter where the binary values for preceding and following pulses are known absolutely, rather than at the receiver where the data values must be deduced.

- The processes of analog data pulse generation and compensation are accomplished in a common circuit function.

There is some detail inconsistency in the following discussion because it has been edited to delete most of the specific material on the impulse response of telephone twisted pairs, however some of the data obtained is generically similar for a bandpass medium. In particular, the twisted pair receiver input filter blocks dc and is a single pole lowpass. This is equal to the most rudimentary radio bandpass function.

GENERAL DESCRIPTION

The technology employed is concerned with optimization of transmitted pulse shape to maximize detectability at the receiving point considering the frequency dependent non-uniformity of receiver bandfiltering in amplitude and phase delay. The inherent shape of spectrum efficient pulse shapes causes intersymbol interference which must be compensated for error-free operation of high rate digital transmissions.

The *a priori* knowledge of the impulse response of interposed band limiting filters is the starting point for the calculation of values used in the compensation network. The development of the desired transmit signal is accomplished in steps by developing the following functions:

1) The development of a received binary line signal which is a composite of overlapping narrow spectrum pulses of binary polarity. This is also the transmitting pulse for a distortionless transmission line.

2) The modification of the selected transmit pulse shape to compensate for bandpass filtering at the receiver to minimize noise susceptibility.

Generation of a Binary Line Signal with Consecutive Overlapping Pulses

The transmitted signal is generated by overlapping pulses of the same or opposite polarities. The pulse shape must have the property of adding to a near constant value when repeated and added at the desired pulse rate.

An illustrative example is shown in Figure 1. A triangular pulse with a width at the base equal to twice the period of the pulse rate has the necessary additive and overlap properties. Information is coded by using one of two polarities for each pulse. The sum of overlapping pulses of the same polarity is a straight line as shown. A further coding step is to define a transition as binary 0 and steady state as binary one.

The chosen shape, described later, is a compressed duobinary pulse spread over two bit intervals (rather than three), which is favorable for transmission in a bandlimited medium and which has the necessary additive property. When the polarity of consecutive pulses is opposite, the line signal makes the transition from one polarity to the other with a near sinusoidal shape.

The pulses are differentially precoded so that the information is carried in the polarity of the current pulse relative to the previous pulse.

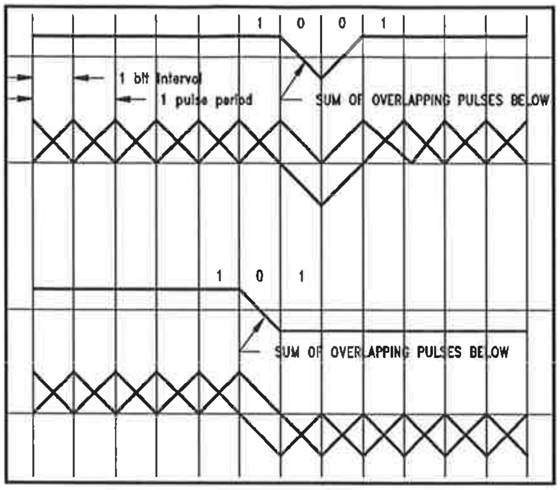


Figure 1--Formation of NRZI Signal with Overlapping Triangular Pulses

Pulse Shape for Distortion Free Medium

Each pulse, optimized for the distortionless medium, is the compressed duobinary pulse given by equation [2] below.

$T^2 \sin(\pi t/T) \;/\; \pi t(T-t)$ [1]
duobinary

$(2T/3)^2 \sin(3\pi t/2T) \;/\; \pi t((2T/3)-t)$ [2]
compressed duobinary

This pulse shape is shown in Figure 4. The width of the main lobe of this pulse is spread over two bit intervals (as given by equation [2] above, rather than the three intervals of well-known duobinary pulses, as given by equation [1].

Consecutive pulses of the same sign add to a near constant dc level provided the detail of the low amplitude parts of the pulse extending a few bit intervals before and after the main lobe are taken into account.

The capacity to define and generate a pulse shape extending over several bit intervals is what makes possible the use of this type of pulse shape many of which must be overlaid in accordance with the data pattern.

This pulse shape is also the desired receiver pulse shape at the point of baseband demodulation after passing through the receiver bandpass filtering.

Characterization of the Medium By Impulse Response

The medium is fully characterized by the impulse response which may be measured or calculated for any specific line model. The impulse response is measured for a transmission line by applying a narrow pulse (narrower than a data pulse) at the send end, and observing the output waveform at the receive end.

> "We now have reached the amazing conclusion that a linear system with constant parameters is uniquely characterized by the single function h(t), which is the system response to an impulse function applied at t=0. It is this function that is termed the system's impulse response to an arbitrary input by the way of the convolution integral."[1]

The impulse response is a measure of the medium. A new function must be generated which undoes the time and amplitude dispersion. This is similar to taking the pulse shown as an input and concentrating it back to the very narrow pulse that was used to generate the shape.

Convolution and Deconvolution

These mathematical arts are described in the literature,[2] where the functions are manipulations using the Fast Fourier Transform. This work can be done with available computer applications, for example:

> **"Convolution and Deconvolution**
> Convolving the impulse response of a digital filter with an input signal generates the corresponding filtered output. This document shows two ways of carrying out convolutions in MathCAD: .... The document also illustrates recovery of the original signal by deconvolution; you can carry out deconvolution efficiently in MathCAD by dividing the FFT of the output sequence by the FFT of the filter impulse response."[3]

DESIGN AND SYNTHESIS OF THE TRANSMIT LINE SIGNAL

The transmit pulse shape is the result of two functions. The first function is the idealized pulse for transmission over a distortionless medium, and the second function is a compensation for the amplitude and time distortion characteristics of the band-limited transmission medium to be used.

The necessary information to compensate transmission distortion is deduced from the impulse response of the medium observed at the receive end of the test transmission line. Also, the impulse response may be calculated or measured on transmission line models.

This process can be considered pre-distorting the transmitted pulse to obtain an optimized result at the receiving point.

The major elements of the process are as described above, and in addition, considerations relating to feasible circuits for generation of the arbitrary analog pulse shape.

Assumptions About Ancillary Functions and Signaling Data Rate

The current design and description is based upon a data throughput of 16 Mbs which, with 4B/5B block coding, is a 20 Mbs line rate and a 10 MHz Nyquist frequency. There is some descriptive ambiguity depending on whether throughput or line rate is most relevant, but no other transmission rates are used in the present description.

Finite Impulse Response (FIR) Pulse Shape Formation

There is existing art[4] on generating certain types of wave shapes using a "finite impulse response" filter. This technique has been used[5] where the pulse shape generated depends on the value of preceding and following bits as well as the current bits. There is a window consideration. In theory, it is necessary to consider all values between + and - infinity, however a judgment must be made as to how wide the time window needs to be for satisfactory results. This value is usually 1 to 3 bits before and after the current bit.

Transmit Pulse Shape

The desired transmit pulse shape is one which, after passing through the distortion introduced by the interposed bandfilters as defined by the impulse response, produces the same pulse shape as was defined for the distortionless case.

The desired transmit pulse shape is calculated as one that a time convolution integral performed on the transmit pulse and the transmission medium impulse response produces the pulse shape defined for the distortion free medium.

This definition starts from the answer and works backward. The method of calculation uses the previously referenced deconvolution procedure.

There is art in deciding how many bits before and after mid-pulse must be considered. If the window is too small, the received "eye" opening will be smaller than it could be. There is circuit and calculation cost in making the window wider than necessary.

As shown in Figure 3, the resultant transmit pulse (and also the weighting function alone) are oscillatory at the data rate. The weighting function may also contain a dc component within a limited time window.

The compensation in this pulse shape is for the low pass and frequency dependent delay of telephone twisted pair. A different oscillatory pulse resembling that shown would be required to offset tight radio band filtering.

RECEIVER BAND-LIMITING

It is desirable to minimize the bandwidth of the receiver input with highpass and lowpass filtering.

The lowpass shape used is a 2-pole near-Butterworth with 1 dB attenuation at the Nyquist frequency which could be characterized as a "loose-fit" filter. The highpass shape is formed by a single series capacitor providing a rolloff of 6 dB per octave. The selected highpass cut-off frequency is the "loose" fit with 100 nfd capacitor. For design of the transmit wave shape, the effect of this filter is included within the network for which the impulse response is obtained and used.

The greater the time delay distorting compensated, the larger the time window (number of bits before and after the current bit) that must be used. Tightening the highpass filter leads to reduction in eye opening as a result of residual dc components.

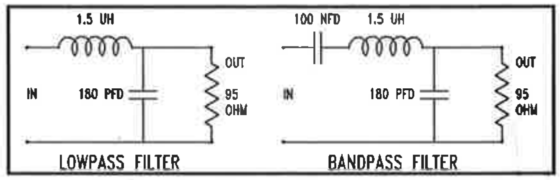


Figure 2--Schematics of Receiver Input Filters

Shown in Figure 2 above are schematic diagrams of the two receiver input filters used in this evaluation and in experiments. Both the high and lowpass are relatively "loosely" fitted to the transmitted spectrum.

Shown below is Figure 3 displaying the transmission characteristics of the bandpass filter obtained by analysis. In both cases, the frequency scale is 1 KHz to 100 MHz.

With this filter added, the calculated transmit pulse and resulting receive pulse is shown in Figures 5 and 6.

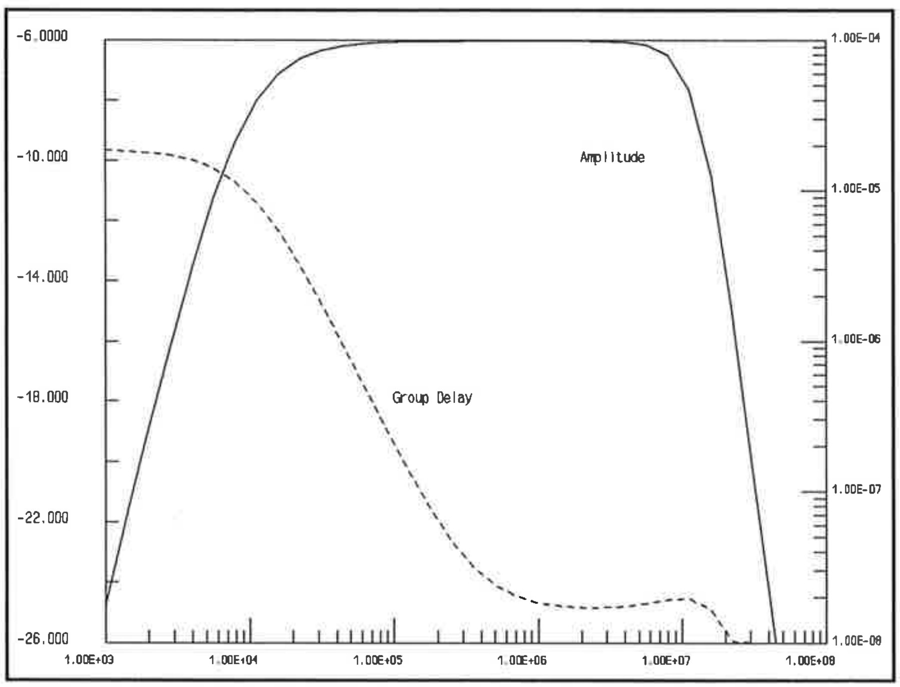


Figure 3--Frequency Response for Bandpass Filter with "Loose" Low Frequency Cutoff from 1 kHz to 100 mHz

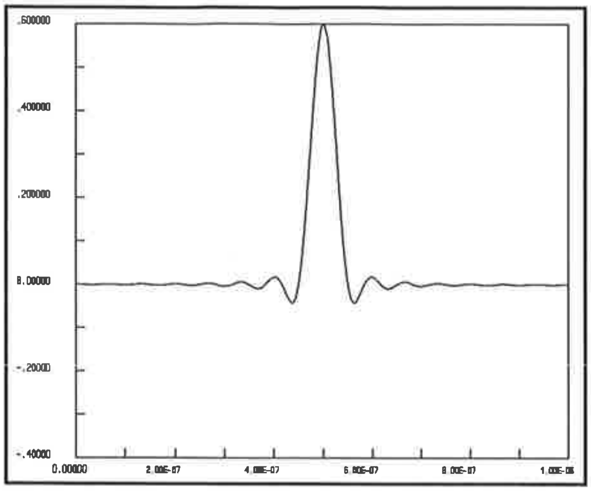


Figure 4--Required Pulse at Input of Demodulator

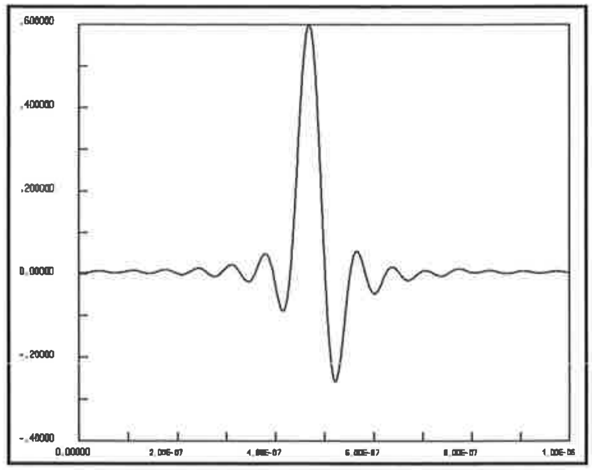


Figure 5--Ideal Transmit Pulse for a Particular Distorted Medium (400 ft DIW)

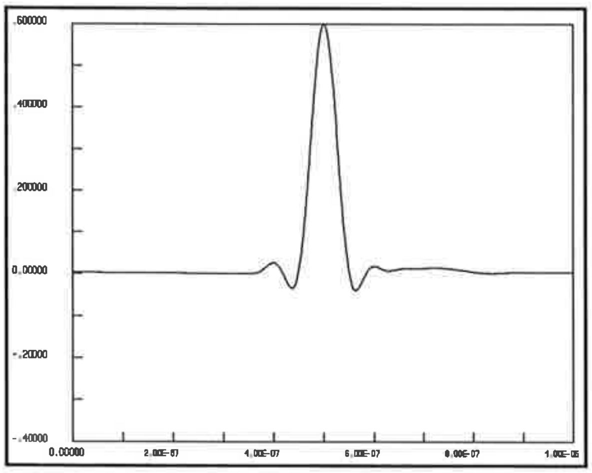


Figure 6--Resulting Receive Pulse Shape After Bandpass Filter with Transmitted Pulse Shape in Figure 5

Note: These Figures relate to 16 Mb/s transmission through 400 feet of DIW, but would be similar for a bandpass radio system. The time scale is 0 to 1000 nanoseconds which is 20 bits at the line rate and 16 bits at the data throughput rate--each division is a 2-bit interval.

REFERENCES:

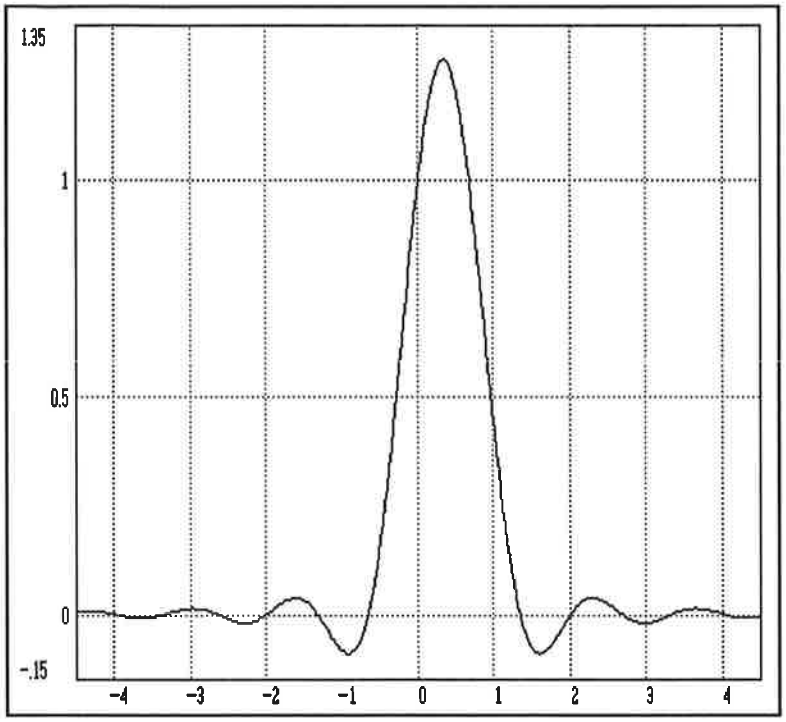
1. "Filtering in the Time and Frequency Domains," H. J. Blinchikoff & Anatol I. Zverev, John Wiley & Sons, New York, 1976, Reprint Edition 1987, ISBN 0-89874-952-2 Page 21

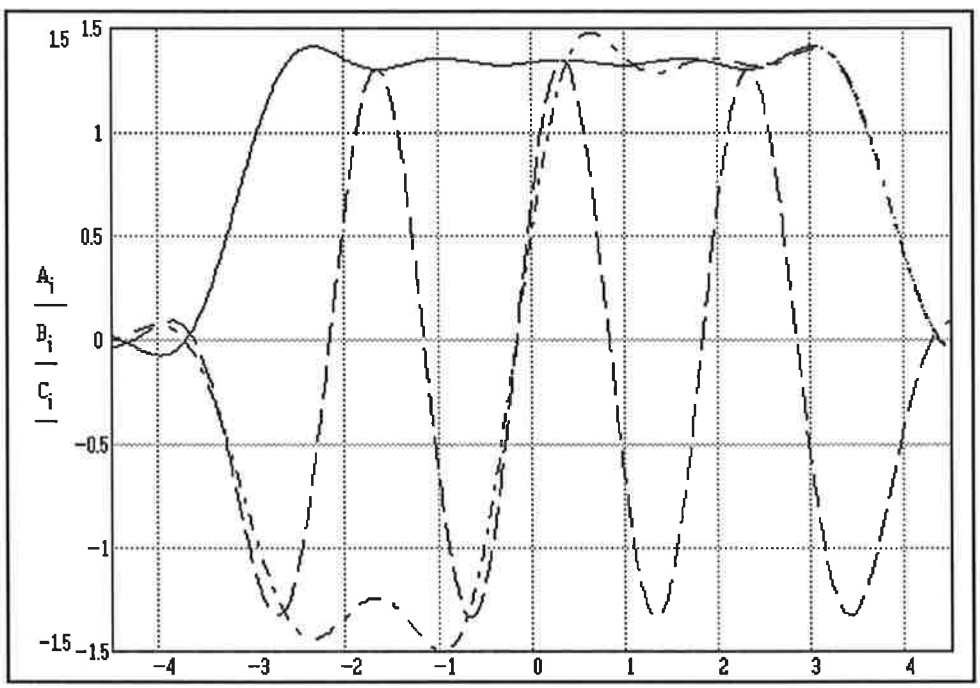
2. "Circuits, Signals and Systems," William McC. Siebert, The MIT Press, Cambridge, 1986

3. "Electrical Engineering Application Pack--MathCAD," Copyright 1989 by MathSoft, Inc., One Kendall Square, Cambridge, MA 02139 (Page 93 of the Manual)

4. "Handbook of Digital Signal Processing Engineering Applications," Edited by Douglas F. Elliott, Academic Press Inc., San Diego, 1987, ISBN 0-12-237075-9

5. U.S. Patent 4,773,082, "RF Modem with Improved Binary Transversal Filter," G. L. Somer





The above figures are the result of calculation and plotting of the 2/3rds width duobinary pulse shape using Mathcad 3.0. The resulting figures were copied into the word processor with some loss of detail.

The upper figure is the isolated pulse, and the lower figure is the sum of seven consecutive pulses with no phase reversals, one phase reversal and alternating phase reversals.