On the application of biphase coding in data communication systems

Eindhoven
University of Technology
the Netherlands

Department of
Electrical Engineering

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SUMMARY

This paper concerns the application of biphase coding for data communication systems where the pulse dispersion is short compared to the bit time of the data signal to be transmitted. As many optical fibre systems have this feature, biphase coding may be very attractive for these systems when the signal-to-noise ratio at the detector input is sufficiently high. In this paper four spectral shaping solutions for biphase coding are worked out. The most satisfactory solution depends upon the design parameters of the system concerned.

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Address of the author:

Ir. A.P. Verlijsdonk,
Telecommunication Division,
Department of Electrical Engineering,
Eindhoven University of Technology,
P.O. Box 513,
5600 MB EINDHOVEN,
The Netherlands

CONTENTS

1. INTRODUCTION

Biphase coding is a special case of mBnB-coding [1,2] with very attractive features with respect to data communication, as there are:

- a constant energy per symbol element,
- much timing information,
- no disparity (i.e. no variation in the DC-component of the coded signal),
- a very simple coder/decoder configuration as Fig. 1 shows.

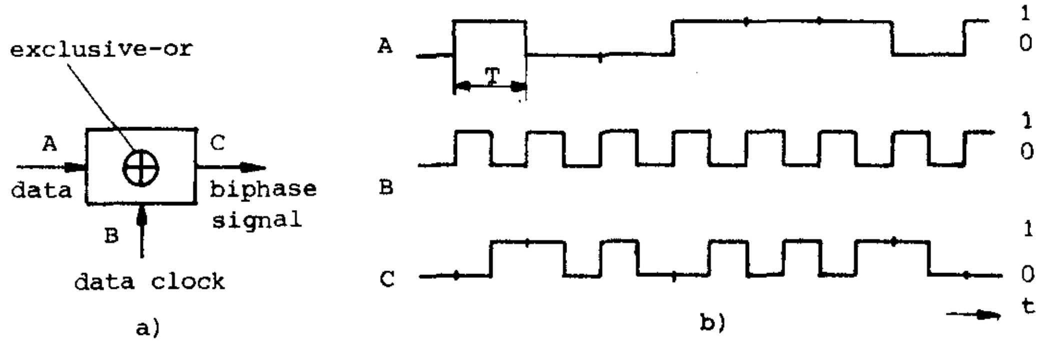

Fig. 1: a) biphase coder, b) signal shapes.

As can be seen from Fig. 1 the coder converts each single data pulse into a double-pulse signal element.
A commonly stated disadvantage of biphase coding is that its bitrate is twice the information rate of the data source; this seemingly corresponds to a poor coding efficiency [3,4].
This interpretation is only valid when both halves of each signal element are separately sampled by the detector.
However, this particular sampling strategy is not required, since the two halves of a double-pulse signal element are completely correlated.
When the biphase signal is regarded as a sequence of double-pulse signal elements it will be obvious that the transmission rate still amounts to $r = \frac{1}{T}$ symbols/s or baud. By means of spectral shaping it is possible to achieve that the eye pattern of the received signal has only one open eye per bittime T. This can be realised by incorporating in the system a linear filter with a transfer characteristic $C(\omega)$ in order that the input signal to the detector satisfies the first Nyquist criterion for a signaling rate of $r = \frac{1}{T}$ symbols/s [5,6].
Fig. 2 shows the system model (sampling at $t = k.T$).

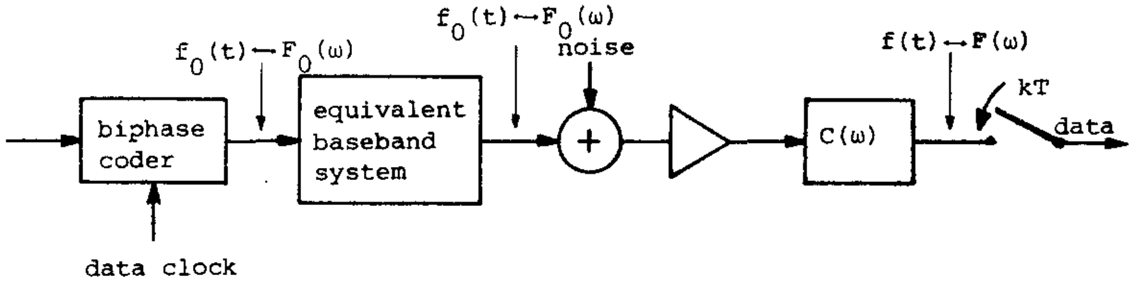

Fig. 2: The system model (sampling at $t = k.T$).

2. THEORETICAL APPROACH

2.1. Sampling at $t = k.T$

The Fourier transform of one biphase signal element $f_0(t)$ is

$$F_0(\omega) = -jT \cdot \frac{\sin^2 \frac{\omega T}{4}}{\frac{\omega T}{4}} \quad , \tag{1}$$

which is imaginary, odd and has zero values for $\omega = k \cdot \frac{4\pi}{T}$, as shown in Fig. 3.

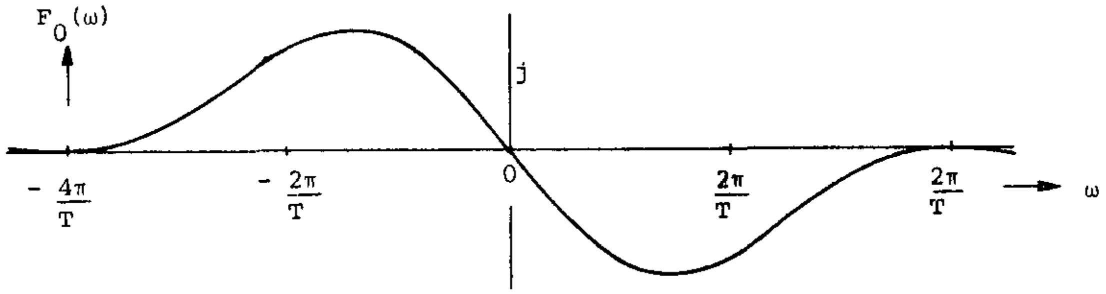

Fig. 3: The Fourier transform of one biphase signal element.

As a consequence it may be clear that there are two main classes of solutions for which $F(\omega)$ can satisfy the first Nyquist criterion for a signalling rate of $r = \frac{1}{T}$ symbols/s.

class 1 (broad-band solutions)

$$F(\omega) = T \quad \text{for} \quad \frac{\pi}{T} < |\omega| \leq \frac{3\pi}{T} \quad \text{and} \quad F(\omega) = 0 \quad \text{elsewhere} \tag{2}$$

or $F(\omega)$ has vestigial symmetry with respect to $|\omega| = \frac{\pi}{T}$ and $|\omega| = \frac{3\pi}{T}$, as indicated in Fig. 4.

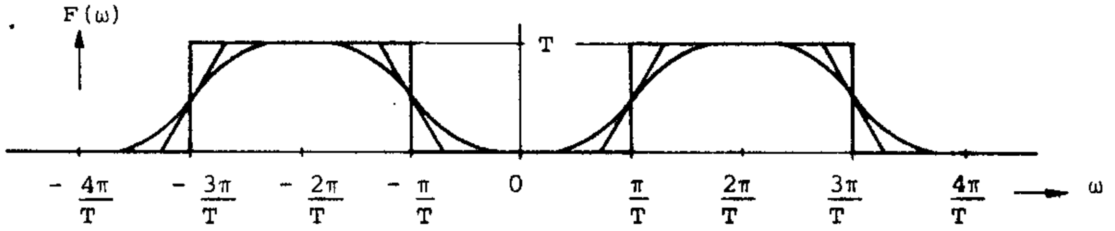

Fig. 4: Class of broad-band solutions for $F(\omega)$ (sampling at $t = k.T$).

Class 2 (narrow-band solutions)

$$F(\omega) = T \quad \text{for} \quad \frac{\pi}{T} \leq |\omega| \leq \frac{2\pi}{T} \quad \text{and} \quad F(\omega) = 0 \quad \text{elsewhere} \tag{3}$$

or $F(\omega)$ has vestigial symmetry around $|\omega| = \frac{\pi}{T}$ and $|\omega| = \frac{2\pi}{T}$, as indicated in Fig. 5.

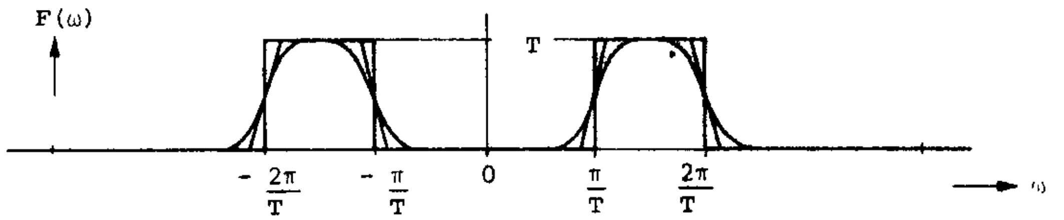

Fig. 5: Class of narrow-band solutions for $F(\omega)$ (sampling at $t = k.T$).

From both classes of solutions the raised-cosine solution is chosen for further investigation.

Particular solution 1 (broad-band; sampling at $t = k.T$)

$$F_1(\omega) = \frac{T}{2}\left(1 - \cos\frac{\omega T}{2}\right) \quad \text{for} \quad 0 \leq |\omega| \leq \frac{4\pi}{T}$$

and (4)

$$F_1(\omega) = 0 \qquad \text{for} \quad |\omega| > \frac{4\pi}{T}$$

as indicated in Fig. 6.

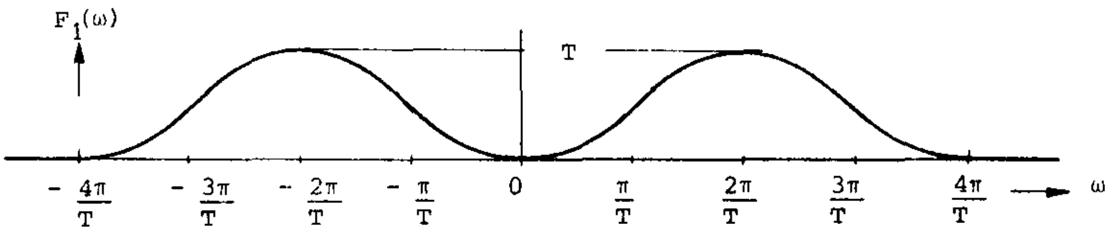

Fig. 6: Particular broad-band solution for $F_1(\omega)$.

For the transfer function $C_1(\omega)$ then follows

$$C_1(\omega) = \frac{\frac{T}{2}\left(1 - \cos\frac{\omega T}{2}\right)}{-jT \cdot \frac{\sin^2\frac{\omega T}{4}}{\frac{\omega T}{4}}} = j\,\frac{\omega T}{4} \qquad \text{for} \quad |\omega| < \frac{4\pi}{T}$$

and (5)

$$C_1(\omega) = 0 \qquad \text{for} \quad |\omega| > \frac{4\pi}{T}$$

as indicated in Fig. 7.

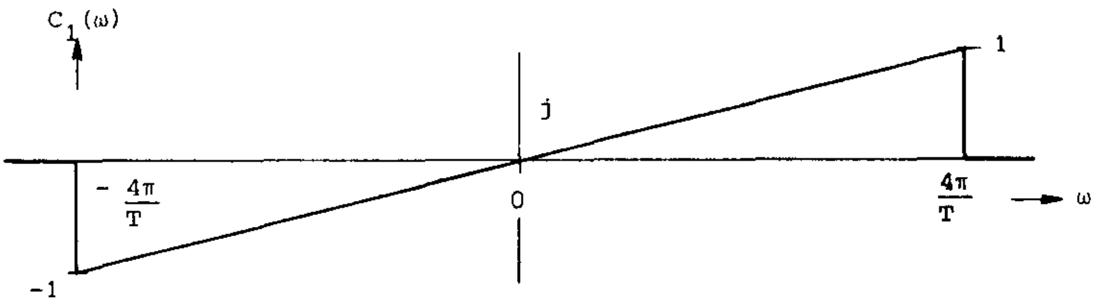

Fig. 7: The transfer function $C_1(\omega)$.

The desired $C_1(\omega)$ can be approximated with the aid of a tapped delay line (TDL), followed by a low-pass filter (LPF) [7]. If the delay time between two successive taps amounts to $\frac{T}{4}$, the transfer function $C_{1D}(\omega)$ of the TDL is periodic with intervals of $\frac{8\pi}{T}$, as indicated in Fig. 8.

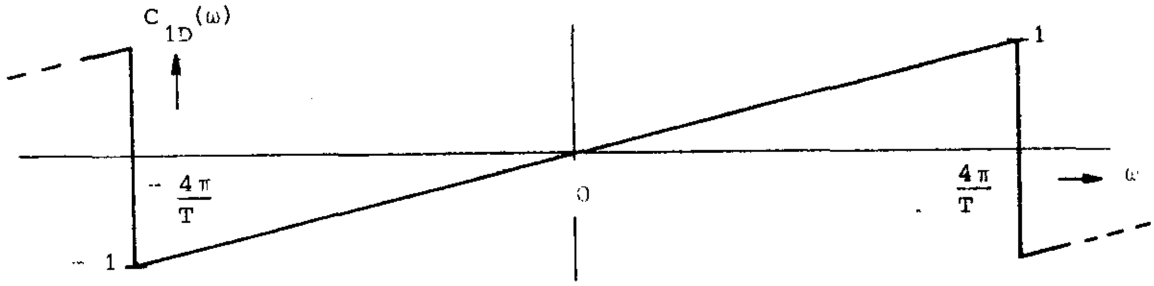

Fig. 8: The periodic transfer function $C_{1D}(\omega)$.

The additional LPF must have a sharp cut-off at the frequency $\omega = \frac{4\pi}{T}$.

Particular solution 2 (narrow-band; sampling at $t = k.T$).

$$F_2(\omega) = \frac{T}{2}(1 - \sin|\omega|T) \quad \text{for} \quad \frac{\pi}{2T} < |\omega| \leq \frac{5\pi}{2T}$$

and (6)

$$F_2(\omega) = 0 \quad \text{for} \quad |\omega| < \frac{\pi}{2T} \quad \text{and} \quad |\omega| > \frac{5\pi}{2T}$$

as indicated in Fig. 9.

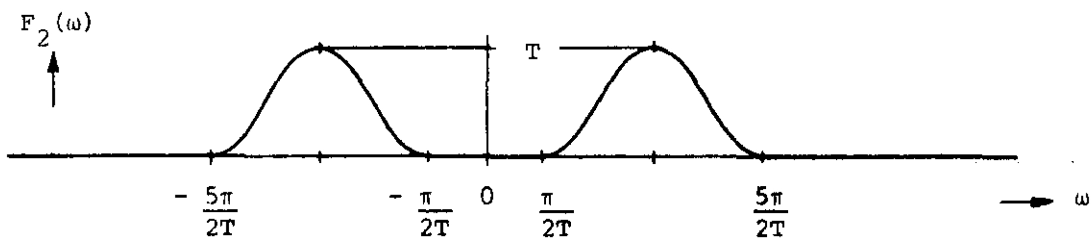

Fig. 9: Particular narrow-band solution $F_2(\omega)$ (sampling at $t = k.T$).

For the desired transfer function $C_2(\omega)$ follows

$$C_2(\omega) = j \cdot \frac{\omega T}{8} \cdot \frac{(1 - \sin|\omega|T)}{\sin^2 \frac{\omega T}{4}} \quad \text{for} \quad \frac{\pi}{2T} \leq |\omega| \leq \frac{5\pi}{2T} \tag{7}$$

and

$$C_2(\omega) = 0 \quad \text{for} \quad |\omega| < \frac{\pi}{2T} \quad \text{and} \quad |\omega| > \frac{5\pi}{2T}$$

as shown in Fig. 10.

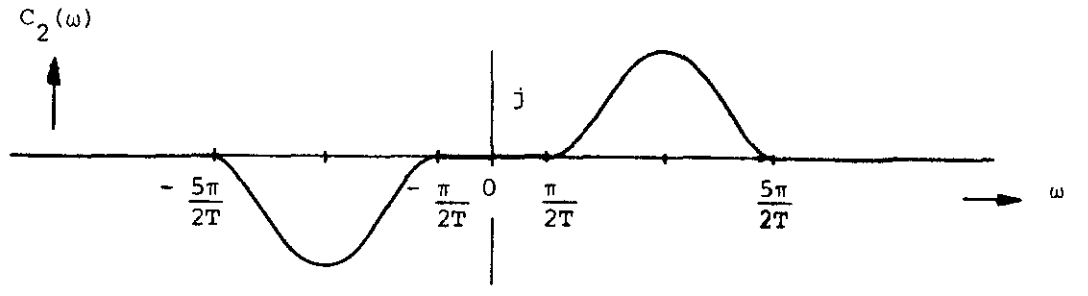

Fig. 10: The transfer function $C_2(\omega)$.

The transfer function $C_2(\omega)$ can also be approximated with the aid of a TDL and a LPF with a cut-off frequency $\omega_g = \frac{5\pi}{2T}$. As there are no strong selectivity requirements, this LPF may have a simple configuration.

As solution 2 implies a smaller signal bandwidth than solution 1, the noise power at the detector input will be smaller than in the case of solution 1.

A further evaluation of both solutions is possible by comparing the corresponding eye openings.

Transformation of $F_1(\omega)$ and $F_2(\omega)$ leads to

$$f_1(t) = \left[2 \cdot \frac{\sin \frac{2\pi t}{T}}{\frac{2\pi t}{T}} + \frac{\sin \left(\frac{2\pi t}{T} + \pi\right)}{\left(\frac{2\pi t}{T} + \pi\right)} + \frac{\sin \left(\frac{2\pi t}{T} - \pi\right)}{\left(\frac{2\pi t}{T} - \pi\right)} \right] \cdot \cos \frac{2\pi t}{T} \quad (8)$$

and

$$f_2(t) = \left[2 \cdot \frac{\sin \frac{\pi t}{T}}{\frac{\pi t}{T}} + \frac{\sin \left(\frac{\pi t}{T} + \pi\right)}{\left(\frac{\pi t}{T} + \pi\right)} + \frac{\sin \left(\frac{\pi t}{T} - \pi\right)}{\left(\frac{\pi t}{T} - \pi\right)} \right] \cdot \cos \frac{3\pi t}{2T} \quad (9)$$

In Fig. 11 five positive and five negative signal elements $f_1(t)$ are drawn.

In Fig. 12 the same is done for $f_2(t)$.

These figures show clearly that for solution 1 the intersymbol interference around the sampling instants is smaller that for solution 2.

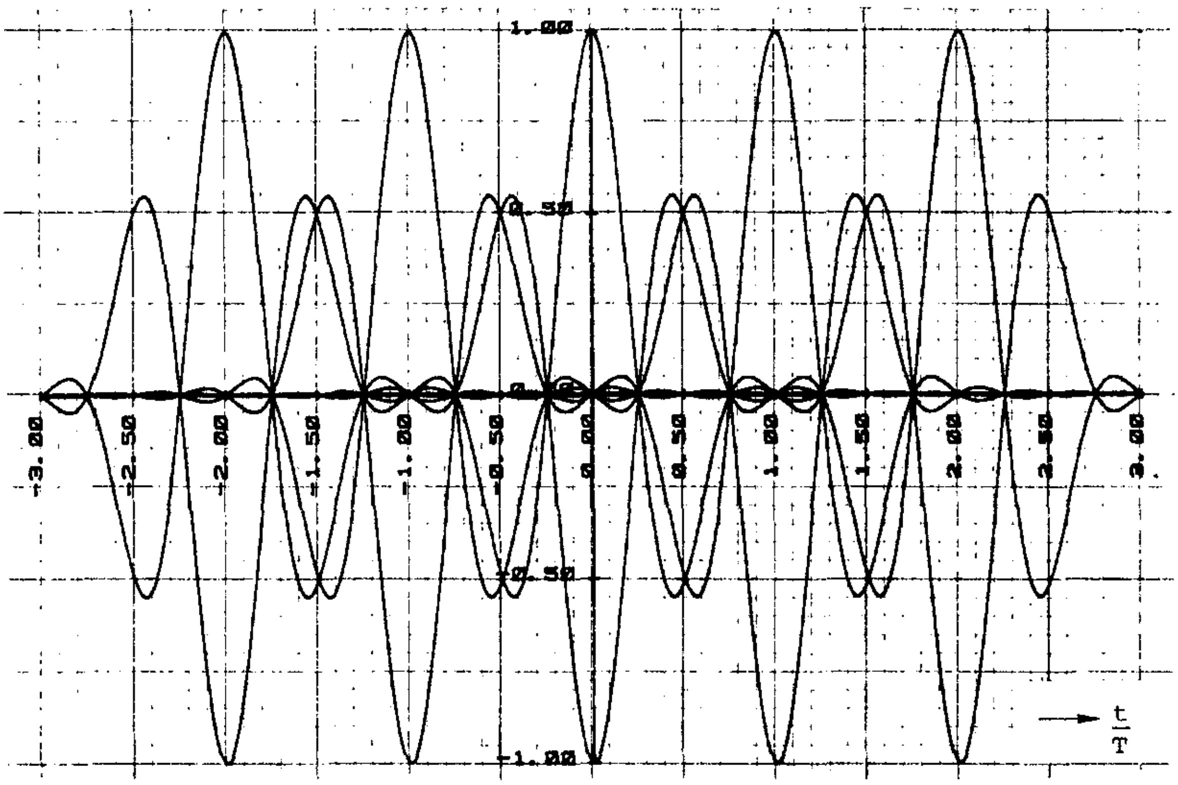

Fig. 11: Five positive and five negative signal elements $f_1(t)$.

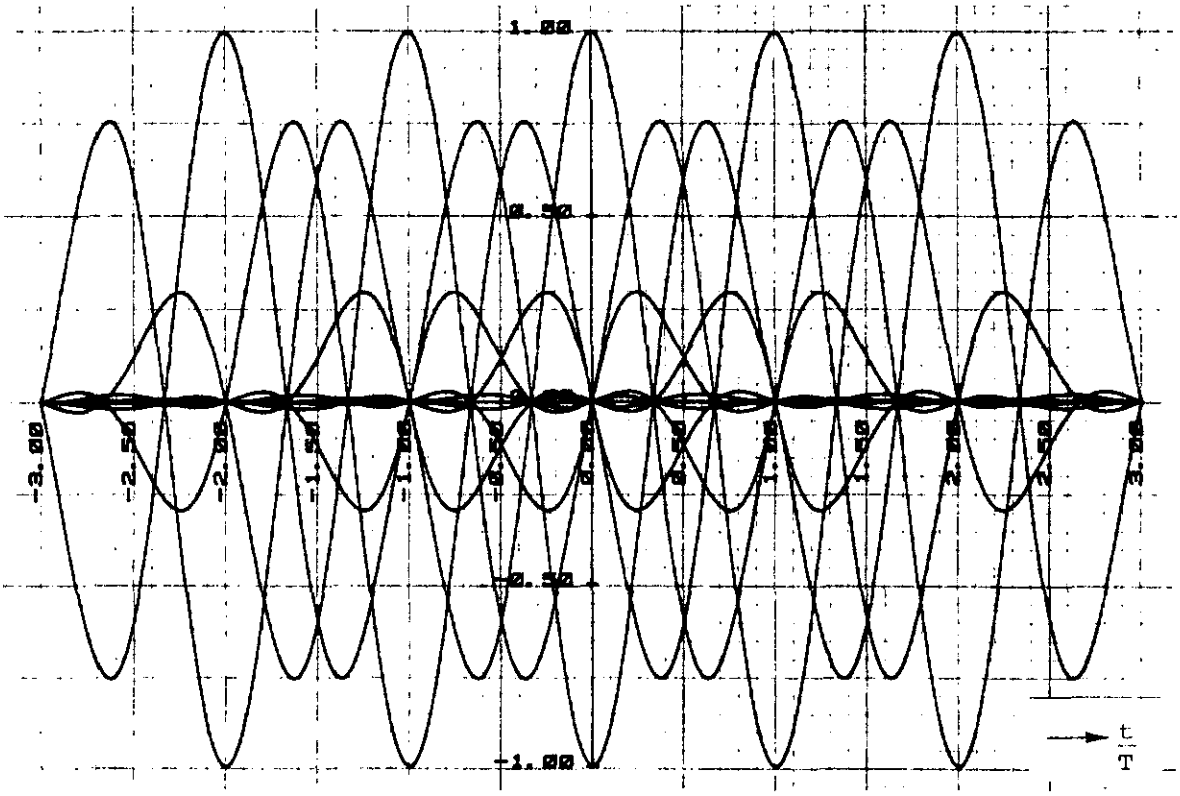

Fig. 12: Five positive and five negative signal elements $f_2(t)$.

The worst case eye opening for both solutions is drawn in Fig. 13. It shows that for solution 1 the horizontal eye opening amounts to $0.5T$, which is larger than $0.3T$ for solution 2.

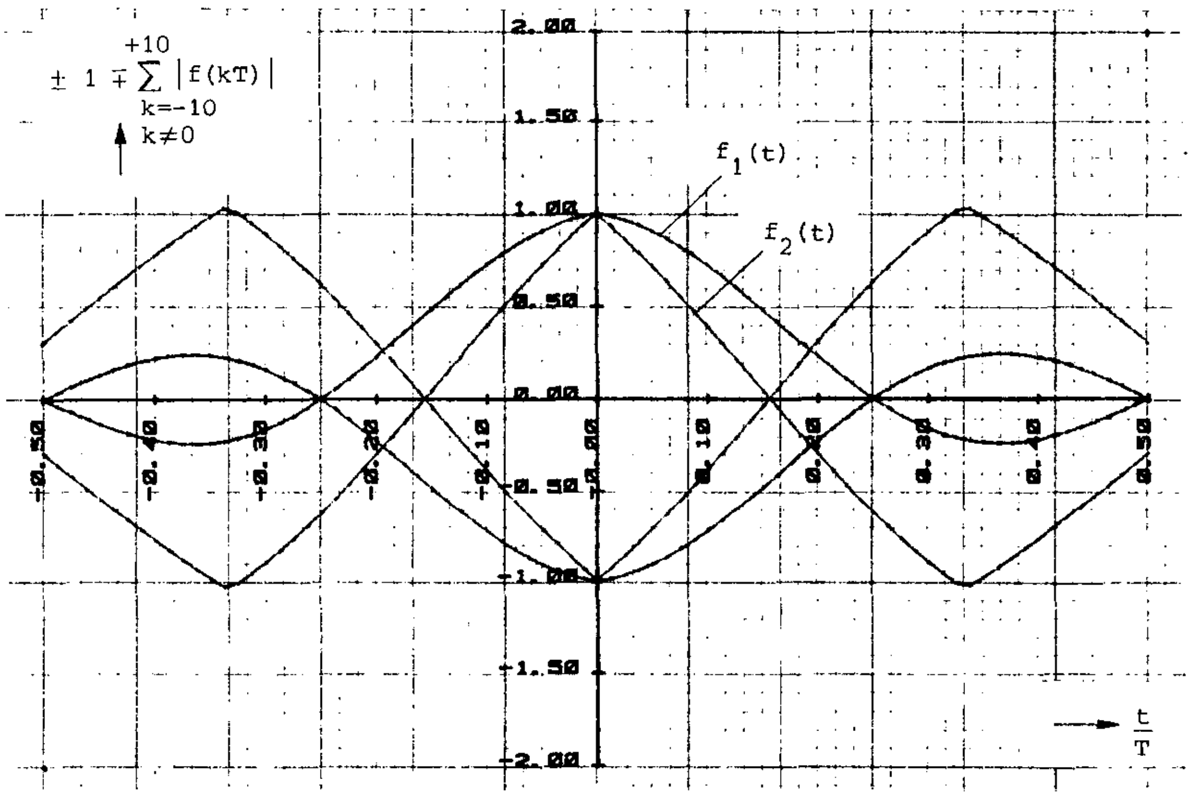

Fig. 13: The worst-case eye openings for $f_1(t)$ and $f_2(t)$.

Intermediate conclusion with respect to the solutions 1 and 2:

In systems where a sufficient signal-to-noise-ratio can hardly be reached, solution 2 will be necessary. When the signal-to-noise-ratio is sufficiently high, solution 1 is preferable, since in that case the detector is less sensitive to clock jitter due to the wider horizontal eye opening.

2.2. Sampling at $t = k.\frac{T}{2}$

In communication systems for which the pulse dispersion is small with respect to the bittime T there is no problem due to the "double bitrate" of a biphase signal. So there is no objection to let the eye pattern of the received signal have two open eyes within one bittime T. The detector may then sample the received signal twice per bittime T as is indicated in Fig. 14.

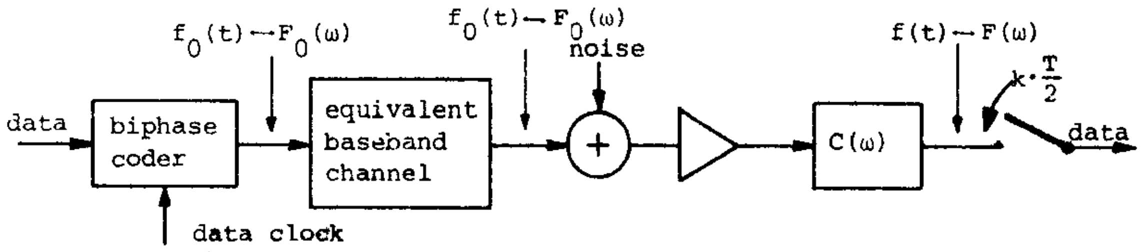

Fig. 14: The system model (sampling at $t = k \cdot \frac{T}{2}$).

In this case the system response $f(t)$ to one signal element should have the following properties:

$$f\left(\frac{2k-1}{2} \cdot \frac{T}{2}\right) = \begin{cases} +1 \text{ for } k = 0 \\ -1 \text{ for } k = 1 \\ \;0 \text{ for } k \neq 0 \text{ and } k \neq 1 \,. \end{cases} \tag{10}$$

These properties can be described by the equation

$$f(t) \cdot \sum_{k + -\infty}^{+\infty} \delta\left(t - \frac{2k-1}{2} \cdot \frac{T}{2}\right) = \delta\left(t + \frac{T}{4}\right) - \delta\left(t - \frac{T}{4}\right) \,. \tag{11}$$

Transforming (11) to the frequency domain yields

$$\sum_{k=-\infty}^{+\infty} F\left(\omega - k \cdot \frac{4\pi}{T}\right) = j.2T.\sin\frac{\omega T}{4} \tag{12}$$

In order to derive the transfer function $C(\omega)$ required to meet this condition, it is useful to consider the simplified system model of Fig. 15.

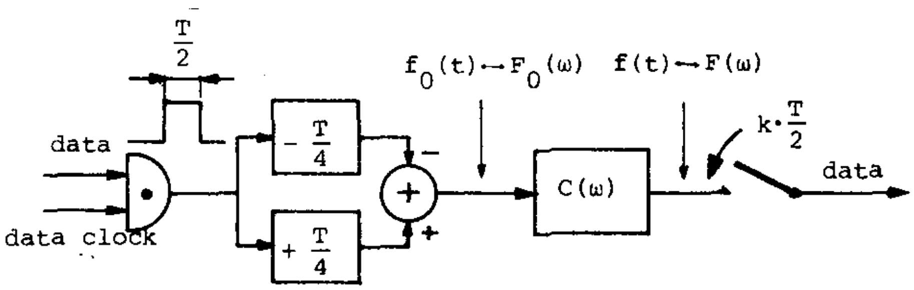

Fig. 15: Simplified system model (sampling at $t = k \cdot \frac{T}{2}$).

From Fig. 15 it will be clear that requirement (10) is fulfilled if the system response $f_h(t)$ to one rectangular pulse of duration $\frac{T}{2}$ satisfies the first Nyquist criterion for a transmission rate of $r = \frac{2}{T}$ baud. This implies that the Fourier transform $F_h(\omega)$ must have vestigial symmetry with respect to $|\omega| = \frac{2\pi}{T}$ [8].

$$F_h(\omega) = \begin{cases} T & \text{for } 0 \leq |\omega| \; \frac{2\pi}{T} - 2\pi\beta, \\ T \cos^2 \frac{1}{8\beta} \left(|\omega| - \frac{2\pi}{T} + 2\pi\beta\right) & \text{for } \frac{2}{T} - 2\pi\beta \leq |\omega| \leq \frac{2\pi}{T} + 2\pi\beta, \\ 0 & \text{for } |\omega| > \frac{2\pi}{T} + 2\pi\beta, \end{cases} \tag{13}$$

as indicated in Fig. 16.

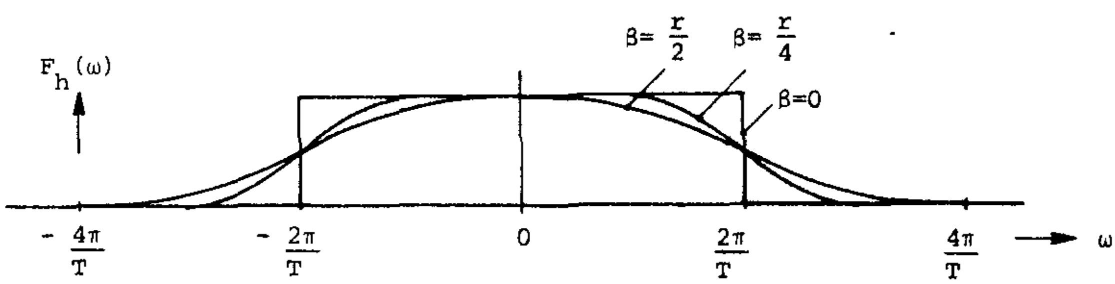

Fig. 16: Nyquist's vestigial symmetry for $r = \frac{2}{T}$ baud.

From Fig,. 15 it follows that

$$F(\omega) = F_h(\omega) \;.\; 2j \sin \frac{\omega T}{4} \tag{14}$$

Out of the class of general solutions for $F_h(\omega)$ are chosen the raised-cosine solution with $\beta = \frac{r}{2} = \frac{1}{T}$ and the narrow-band solution with $\beta = 0$ (theoretical limit). These are nominated as particular solutions 3 and 4, respectively.

Particular solution 3 (broad-band, sampling at $t = k \cdot \frac{T}{2}$).

$$F_3(\omega) = F_{3h} \cdot 2j \sin \frac{\omega T}{2}$$

$$= \frac{T}{2} \left(1 + \cos \frac{\omega T}{4}\right) \cdot 2j \sin \frac{\omega T}{4} \quad \text{for} \quad 0 < |\omega| < \frac{4\pi}{T} \qquad (15)$$

and $F_3(\omega) = 0$ for $|\omega| > \frac{4\pi}{T}$,

as indicated in Fig. 17.

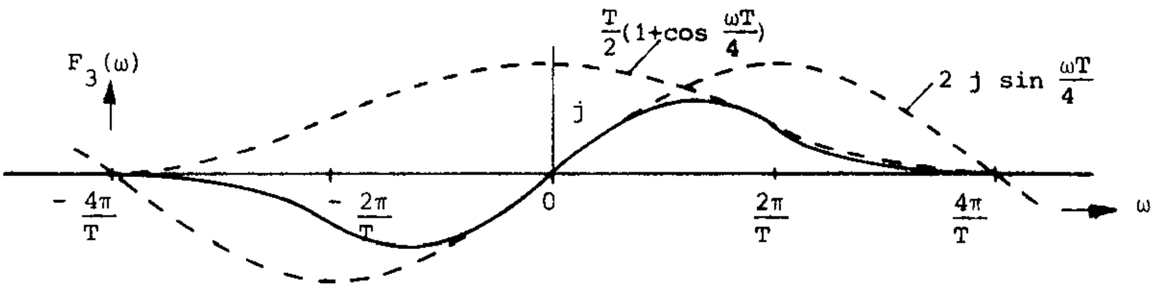

Fig. 17: Particular broad-band solution $F_3(\omega)$ (sampling at $t = k \cdot \frac{T}{2}$).

For the transfer function $C_3(\omega)$ then follows

$$C_3(\omega) = \frac{\frac{T}{2}\left(1 + \cos \frac{\omega T}{4}\right)}{T \frac{\sin \frac{\omega T}{4}}{\frac{\omega T}{4}}} = \frac{\omega T}{8} \cdot \operatorname{cotan} \frac{\omega T}{8} \quad \text{for} \quad 0 \leq |\omega| \leq \frac{4\pi}{T} \qquad (16)$$

and

$$C_3(\omega) = 0 \quad \text{for} \quad |\omega| > \frac{4\pi}{T},$$

as shown in Fig. 18.

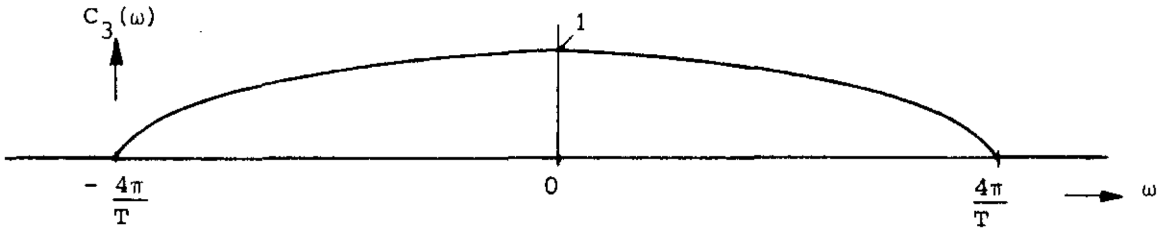

Fig. 18: The transfer function $C_3(\omega)$.

The transfer function $C_3(\omega)$ can also be approximated with the aid of a cascade of a TDL and a LFP. If the delay time between two successive taps amounts to $\frac{T}{4}$ then the transfer function $C_{3D}(\omega)$ of the TDL is periodic with intervals of $\frac{8\pi}{T}$. The TDL must have many taps to approximate the desired $C_{3D}(\omega)$ and the LPF must have a sharp cut-off at the angular frequency $\omega = \frac{4\pi}{T}$.

Particular solution 4 (narrow-band, sampling at $t = k \cdot \frac{T}{2}$)

$$F_4(\omega) = F_{4h} \cdot 2j \sin \frac{\omega T}{4}$$

$$= T \cdot 2j \sin \frac{\omega T}{2} \qquad \text{for} \quad 0 \leq |\omega| \leq \frac{2\pi}{T} \tag{17}$$

and $F_4(\omega) = 0$ for $|\omega| > \frac{2\pi}{T}$,

as indicated in Fig. 19.

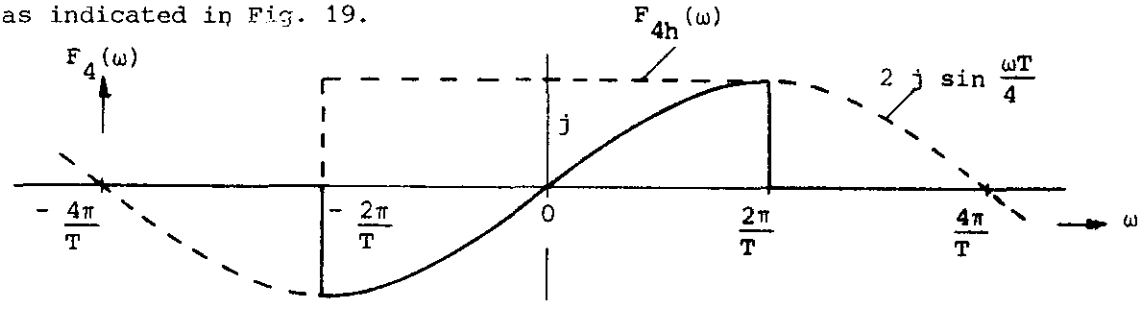

Fig. 19: Particular narrow-band solution 4 (sampling at $t = k \cdot \frac{T}{2}$).

For the transfer function $C_4(\omega)$ then follows

$$C_4(\omega) = \frac{\frac{\omega T}{4}}{\sin \frac{\omega T}{4}} \quad \text{for} \quad 0 \leq |\omega| \leq \frac{2\pi}{T} \tag{18}$$

and

$$C_4(\omega) = 0 \qquad \text{for} \quad |\omega| > \frac{2\pi}{T} ,$$

as indicated in Fig. 20.

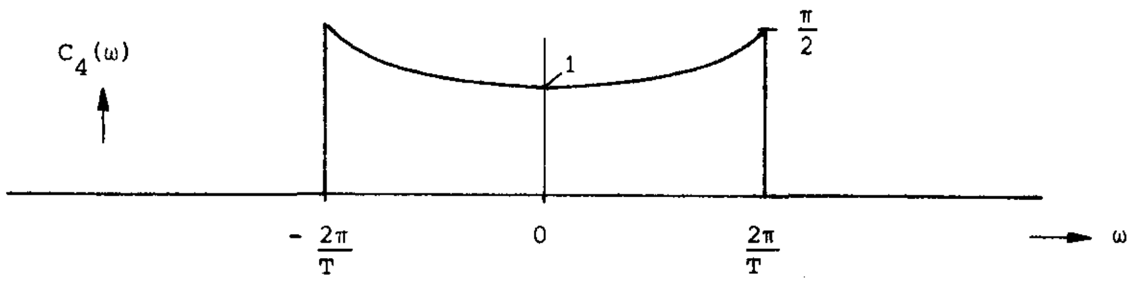

Fig. 20: The transfer function $C_4(\omega)$.

This transfer function $C_4(\omega)$ can also be approximated with the aid of a TDL followed by a LPF. For a delay time of $\frac{T}{4}$ between two successive taps the transfer function $C_{4D}(\omega)$ of the TDL will be periodic with intervals $\frac{8\pi}{T}$. The TDL must have many taps to approximate the desired $C_{4D}(\omega)$ but the LPF may have a simple configuration as there are no strong selectivity requirements.

Further evaluation of the solutions 3 and 4 is done in this time domain by comparing the corresponding eye openings.
Transformation of $F_3(\omega)$ and $F_4(\omega)$ leads to

$$f_3(t) = \frac{\sin\left(\frac{4\pi t}{T} + \pi\right)}{\left(\frac{4\pi t}{T} + \pi\right)} - \frac{\sin\left(\frac{4\pi t}{T} - \pi\right)}{\left(\frac{4\pi t}{T} - \pi\right)} +$$

$$+ \frac{1}{2} \cdot \frac{\sin\left(\frac{4\pi t}{T} + 2\pi\right)}{\left(\frac{4\pi t}{T} + 2\pi\right)} - \frac{1}{2} \cdot \frac{\sin\left(\frac{4\pi t}{T} - 2\pi\right)}{\left(\frac{4\pi t}{T} - 2\pi\right)} \tag{19}$$

and

$$f_4(t) = \frac{\sin\left(\frac{2\pi t}{T} + \frac{\pi}{2}\right)}{\left(\frac{2\pi t}{T} + \frac{\pi}{2}\right)} - \frac{\sin\left(\frac{2\pi t}{T} - \frac{\pi}{2}\right)}{\left(\frac{2\pi t}{T} - \frac{\pi}{2}\right)} \qquad (20)$$

In Fig. 21 five positive and five negative signal elements $f_3(t)$ are drawn.

In Fig. 22 the same is done for $f_4(t)$.

These figures show clearly that the intersymbol interference near the sampling instants is smaller for solution 3 than for solution 4.

The worst-case eye opening for both solutions is drawn in Fig. 23. It shows that for solution 3 there are two eye openings within one bittime T, each being $\frac{t}{2}$ s. For solution 4 the horizontal eye opening is very small. In fact it tends to zero when the number of interferers tends to infinity. The eye openings for solution 4 are drawn twice; once for $N = 10$ preceeding and $N = 10$ following interfering responses and once for $N = 40$ preceeding and $N = 40$ following interfering responses.

Intermediate conclusion with respect to solutions 3 and 4:

In systems where a sufficient signal-to-noise ratio can hardly be realised, a solution near that of solution 4 will be necessary. If the signal-to-noise ratio is sufficiently high, solution 3 will be preferable since in that case the detector is less sensitive to clock jitter due to the wider horizontal eye opening. A particular solution with $\beta = \frac{r}{4} = \frac{1}{2T}$ would be a reasonable compromise. This solution would also be a comprise between the TDL en LPF complexity.

3. Implementation of the shaping network

As mentioned before the shaping network can be implemented by a TDL followed by a LPF. However, a TDL is very difficult to implement at the receiver side, but, when placed after the biphase coder it can be implemented as a binary transversal filter. Even the biphase coder function can be performed by this binary transversal filter. The LPF should stay at the receiver side to restrict the receiver noise. To eliminate unwanted intermodulation products an extra LPF in the transmitter may be useful. Fig. 24 shows the practical system configuration.

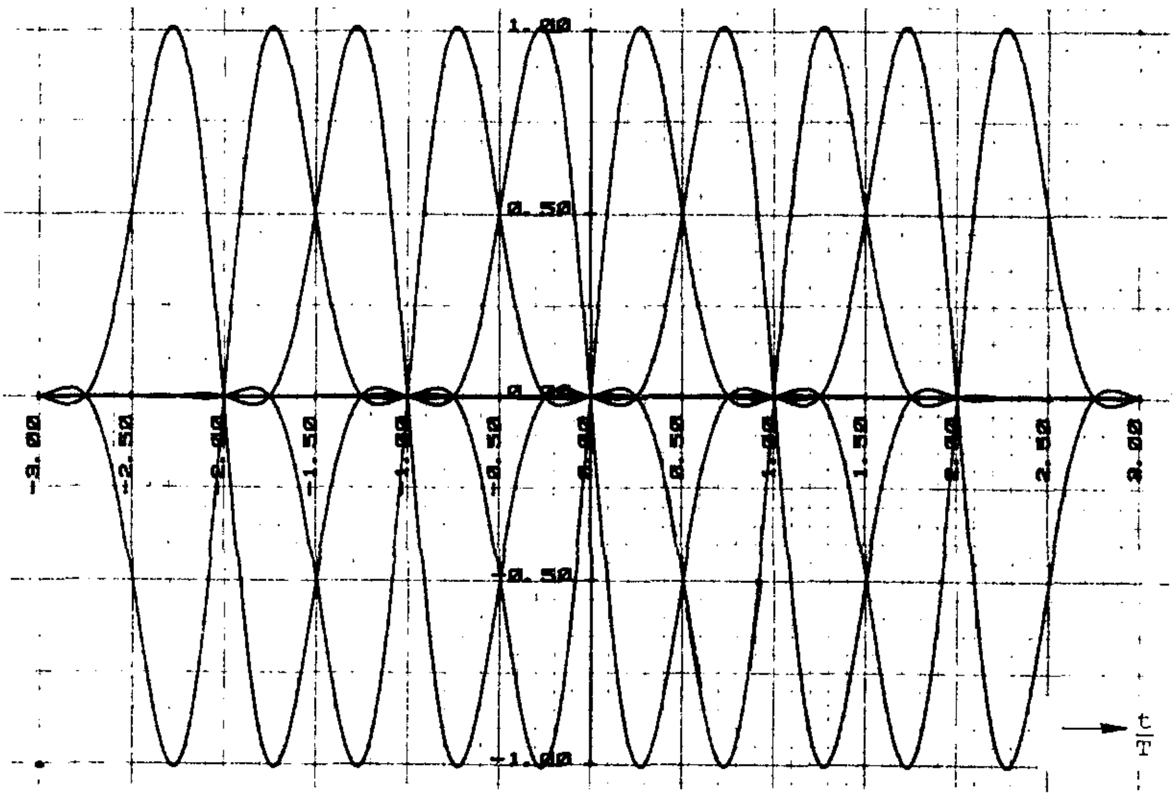

Fig. 21: Five positive and five negative signal elements $f_3(t)$.

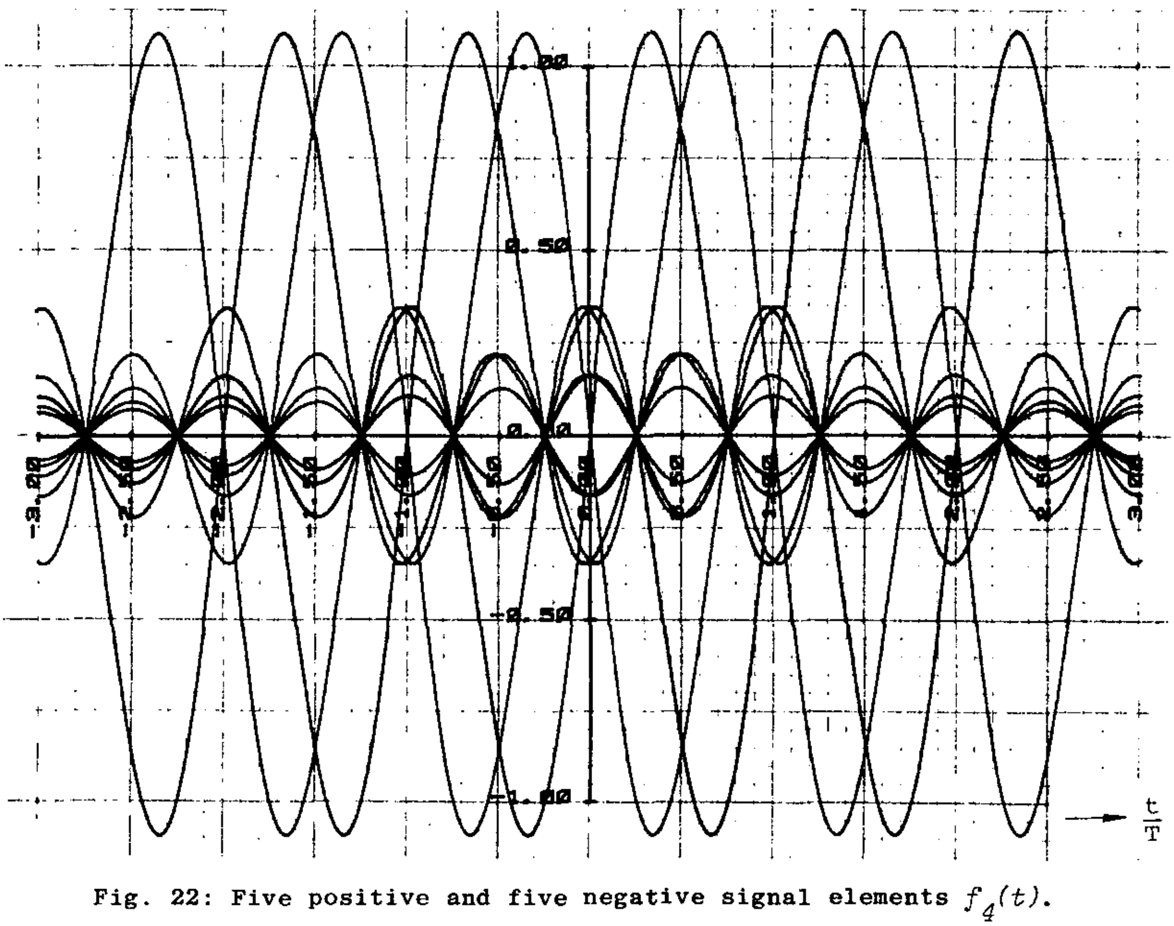

Fig. 22: Five positive and five negative signal elements $f_4(t)$.

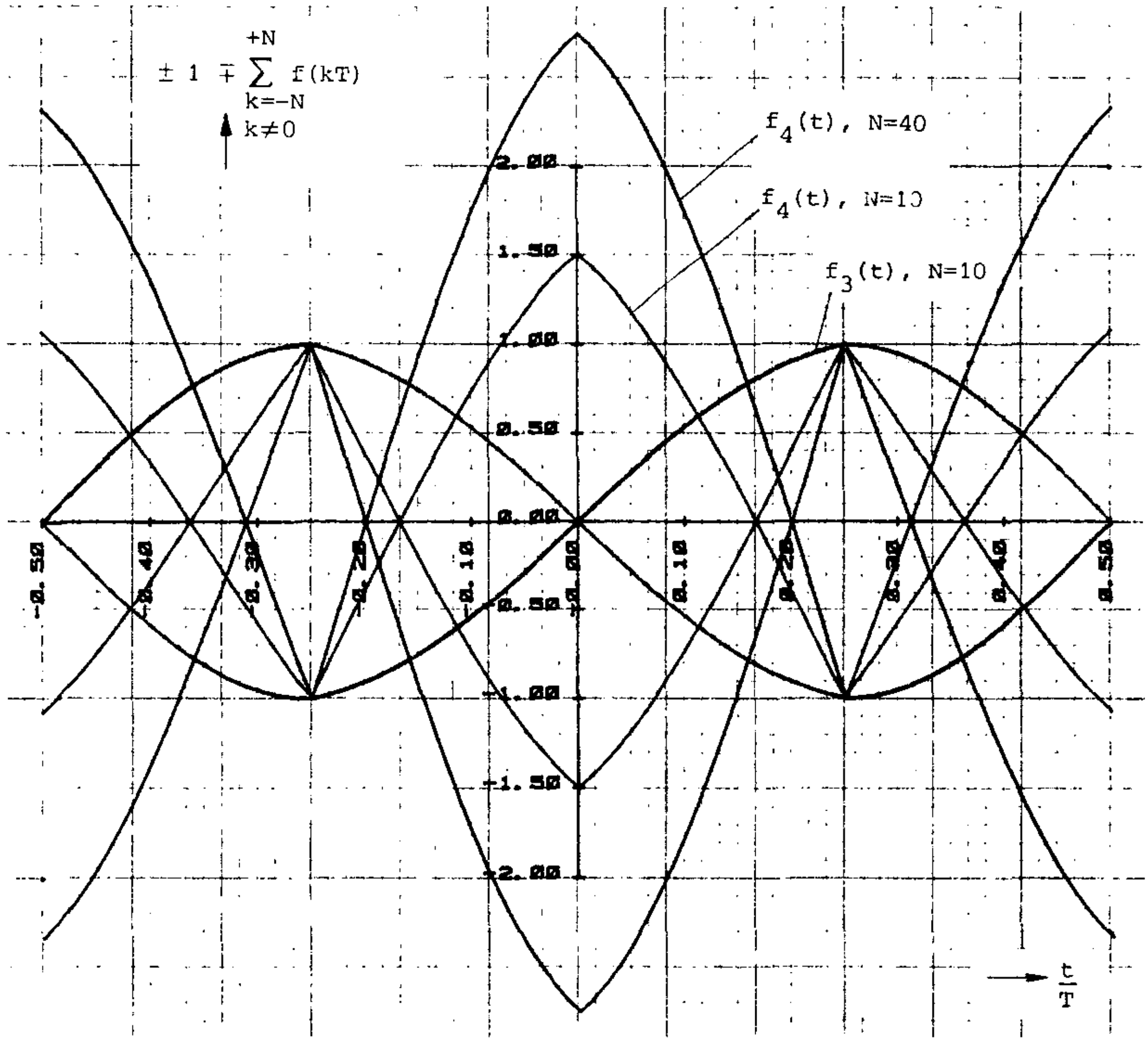

Fig. 23: The worst-case eye openings for $f_3(t)$ and $f_4(t)$.

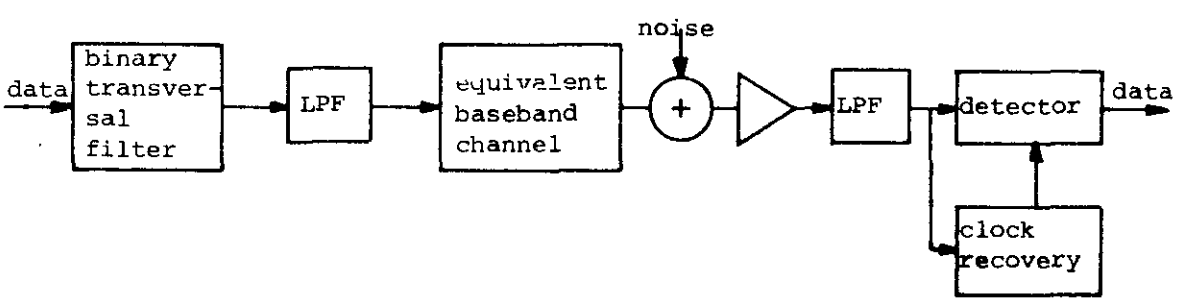

Fig. 24: The practical system configuration.

4. Data clock recovery

For the particular solutions 1 and 3 the data clock can be recovered by detecting the periodic zero crossings of the received signal. By differentiation and full-wave-rectification of the clipped signal a frequency component of $\frac{2}{T}$ Hz is generated, which can be filtered out by means of a Phase Locked Loop.
After dividing the frequency by a factor of 2 the data clock is available with correct frequency but with an ambiguity in the phase, due to the starting position of the frequency dividing flip-flop.
For solution 1 an automatic reset circuit should be built in to prevent the data clock from having the wrong phase.
For solution 3 such a reset provision is not necessary as a wrong phase position causes an inversion of the data signal polarity. The effects of this can be eliminated by the application of differential coding on the original data signal and differential decoding of the recovered data signal.
For solutions 2 and 4 clock recovery is possible by detection of the peaks in the received signal; this method is slightly more complicated and will give rise to more clock jitter.

5. Survey and conclusions

The following table gives a survey of the different solutions

	Sampling once/bittime T		Sampling twice/bittime T	
	Solution 1	Solution 2	Solution 3	Solution 4
bandwidth in Hz	$\frac{2}{T}$	$\frac{5}{4} \cdot \frac{1}{T}$	$\frac{2}{T}$	$\frac{1}{T}$
clock recovery	rather simple	more complicated	very simple	more complicated
clock jitter	low	rather low	low	rather low
number of taps TDL	many	rather many	many	rather many
selectivity LPF	high	low	high	low

The application of biphase coding is very attractive in data communication systems with a good signal-to-noise ratio and a low pulse dispersion compared to the bittime T of the transmitted data signal. Most

optical fibre systems have these features.
Application of biphase coding is especially attractive for optical fibre systems as it permits stable biasing of the laser diode and allows AC-coupling of the front-end amplifier to the photo-diode.
For other transmission systems the application of biphase coding may also be advantageous, but careful signal shaping will be necessary to attain a sufficiently low bit error rate.
Which particular solution should be chosen depends upon the system design parameters, but solutions 1 and 3 look very elegant.

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