

(19) World Intellectual Property Organization
International Bureau



(43) International Publication Date
26 May 2006 (26.05.2006)

PCT

(10) International Publication Number
WO 2006/054131 A1

(51) International Patent Classification:
H04L 27/22 (2006.01)

(21) International Application Number:
PCT/IB2004/004416

(22) International Filing Date:
19 November 2004 (19.11.2004)

(25) Filing Language: English

(26) Publication Language: English

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(81) Designated States (unless otherwise indicated, for every kind of national protection available): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NA, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, YU, ZA, ZM, ZW.

(84) Designated States (unless otherwise indicated, for every kind of regional protection available): ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IS, IT, LU, MC, NL, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Declaration under Rule 4.17:

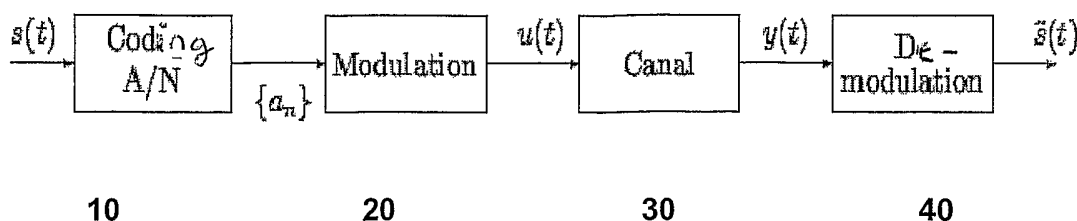
— of inventorship (Rule 4.17(iv))

Published:

— with international search report

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: METHOD FOR DEMODULATING CPM SIGNALS



(57) Abstract: The invention concerns a method for demodulating a received CPM signal ($y(t)$) having a sampling period T and a carrier frequency (w_c) and received from a communication channel, characterized in that it comprises the following steps carried out in real time as said received CPM signal is received: Said received CPM signal ($y(t)$) is received as a continuous signal. Without any discretization of said received CPM signal along the sampling period, this signal is multiplied to yield an inphase component and a quadrature phase component, for each successive time interval during which a symbol a_n is transmitted, carry out the following operations: Define respective first signals for said inphase component and for said quadrature phase component, define second signals $p(\tau)$ and $q(\tau)$ such that the derivative of said first signals with respect to τ satisfy a differential equation, treat said differential equation in order to obtain a linear equation, and retrieve for each time interval where a CPM signal is received a corresponding symbol a_n .

Method for demodulating CPM signals

The present invention concerns the demodulation of CPM signals.

More precisely, the invention concerns a method for demodulating
5 CPM signals, and a method for transmitting a signal with a CPM
modulation/demodulation. The invention also concerns an associated
apparatus.

CPM ("Continuous Phase Modulation") refers to a type of technique
for modulating a digital signal (said digital signal can correspond to the
10 output of a A/D converter fed at its input by an analogical signal). A
description of this type of technique can be found e.g. in "Continuous phase
modulation – Part I and II", IEEE Trans. on Comm., vol. 29, n° 3. pp. 196-
225, 1981.

Signals modulated by such a CPM technique (CPM signals) indeed
15 present a constant envelope and a continuously varying phase. These two
features make the CPM both power and bandwidth efficient and thus, very
attractive for applications such as wireless communication systems.

As an example, the Gaussian Minimum Shift Keying (GMSK)
modulator used in the European GSM wireless system employs such a
20 CPM technique.

Figure 1 shows the general scheme of a digital transmission system
used for the transmission and exploitation of signals modulated by a CPM
technique.

In such system, a continuous incoming signal $s(t)$ is first sampled by
25 an unit 10 with a sampling frequency $F_e = 1/T$, where T is the sampling
period. The resulting stream of binary data which forms the output of unit 10
is then separated into packets of N bits and coded into a series of symbols
 $\{a_n\}$, said symbols being selected from an register having $M = 2^N$ entries.

A CPM modulator 20 then codes the series of symbols $\{a_n\}$ into a
30 continuous signal $u(t)$ which can be transmitted over a communication
channel 30. Said channel can be any type of communication channel, e.g. a
wireless channel.

More precisely, the output $y(t)$ of the channel 30 is not identical to its input $u(t)$.

The channel 30 indeed corrupts the input signal with :

- an intersymbol interference (which corresponds to a linear filtering of the signal to be transmitted – such filtering is modeled by a convolution $f(t)*u(t)$ of $u(t)$ by the impulse response $f(t)$ of the channel), and with
- an additive noise $b(t)$, which generally corresponds to a high frequency component (this noise component represents a perturbation which varies quickly in time when compared to the other elements of the CPM signal).

Thus $y(t)=f(t)*u(t)+b(t)$ where $f(t)$ represents the impulse response of the channel and where the symbol $*$ denotes the convolution operator. The intersymbol interference is due to the impulse response $f(t)$.

The channel 30 allows the communication between an emitter, which includes (or is in communication with) elements 10 and 20, and a receiver, which includes (or is in communication with) a demodulator 40.

The demodulator 40 receives the signal $y(t)$ and retrieves the symbols a_n through a demodulation method.

More precisely, the existing demodulation methods either :

- comprise a preliminary step which accounts to compensate for the perturbation effect of the impulse response of the channel. In such case the channel is “equalized” by a linear filter called the equalizer. The output of the said equalizer, with input $y(t)$, corresponds to an equalized received signal, say $r(t)$.
 - In such case the equalizer is assumed to be “perfect”, and the equalized received signal takes the form : $r(t) = u(t) + v(t)$, where $v(t)$ is a noise corruption,
 - It is to be noted that such preliminary step requires significant computational resources (especially for CPM signals),
- or merely neglect the perturbation effect of the impulse response of the channel – among others to avoid complex and heavy calculations.
 - In such case , the convolutive effect of the channel is neglected,
 - And the received signal is assumed to be of the form $u(t) + b(t)$.

In both cases, the signal to be demodulated is assumed to be of the general form $(u(t) + \text{noise})$, with the following :

- If the perturbation effect of the impulse response of the channel has been taken into account :
 - 5 ➤ the equalizer is assumed to be perfect (which is not the case), and
 - heavy and complex calculations are required,
- alternatively, if the perturbation effect of the impulse response of the channel has been ignored :
 - 10 ➤ the assumed form of the signal to be demodulated is different from the actual received signal $y(t)$ and the demodulation process is likely to be highly inaccurate.

Next, the noisy modulated signal $r(t)$ is first sampled in order to build a discretized representation of said signal, and this discretized
15 representation is then processed in order to carry out the demodulation.

These known methods are generally based on an optimization scheme which uses e.g. a correlator followed by a maximum-likelihood estimator using a Viterbi-like algorithm to search for the minimum of some distance path, through a state trellis. In such case the demodulation method
20 shall be referred to as an "optimal" method (in the minimum mean square sense).

Optimal methods are however associated with an important limitation: they imply complex calculations which require very high computational resources.

25 Alternatively, other known methods use a bench of matched filters for symbol-per-symbol detection. These methods are referred to as "sub-optimal" methods.

Sub-optimal methods allow a more simple processing for demodulating the signal received from the communication channel.
30 However, these methods still correspond to a high computational burden.

Furthermore, the known methods mentioned above - optimal or sub-optimal - are all based on a "probabilistic" approach.

These methods thus require a prior knowledge of the distribution for the symbols $\{a_n\}$, as well as the distribution of the disturbance corresponding to the noise $b(t)$.

And these known methods are - as mentioned above – associated
5 to high computational burdens (which tends to generate additional delays in the demodulation process), and/or do not take into account the perturbation effect of the impulse response of the channel.

These known methods are thus associated to significant limitations.

It is an object of the invention to overcome these limitations.

10 And generally, it is an object of the invention to allow demodulating CPM signals in real-time, in a streamlined and continuous manner, without requiring as large computational resources as the known methods and without ignoring the convolution effect of the channel.

Another object of the invention is to eliminate (or at least reduce in a
15 very significant manner) the influence of the “noise” component $b(t)$ mentioned above.

In order to attain these objects, the invention proposes a method for demodulating a received CPM signal having a sampling period T and a carrier frequency (ω_c) and received from a communication channel, said
20 received CPM signal being of the general form :

$$y(t) = A \int_{-\infty}^{\infty} f(\lambda) \cos \left(\omega_c(t-\lambda) + \pi h a_n \left[\frac{t-\lambda}{T} - n \right] + \pi h \sum_{k=-\infty}^{n-1} a_k \right) d\lambda + b(t)$$

where the convolution by the impulse response of the channel $f(t)$ represents the perturbation effect of the channel, and where $b(t)$ is an additive noise; for the n th sampling interval of the signal :

25 $t = \text{time} = (n + \tau)T$, $0 \leq \tau \leq 1$, h : modulation index, and $\varphi_n = \pi h \sum_{k=0}^{n-1} a_k - n\omega_c T$,

said received CPM signal ($y(t)$) being obtained from :

➤ an original modulated signal ($u(t)$) generated from a series of symbols (a_n), said series being itself generated from an initial signal to be

transmitted via said communication channel, said original modulated

signal being of the general form for $nT \leq t \leq (n+1)T$:

$$u(t) = A \cos(\omega_n t + \varphi_n) \quad \tau = t/T - n$$

(equation(1)).

with : $\omega_n = \omega_c T + \pi h a_n$, and

- 5 ➤ a signal (b(t)) corresponding to a noise generated by the transmission by said communication channel,

said method being characterized in that it comprises the following steps carried out in real time as said received CPM signal is received :

- Said received CPM signal is received as a continuous signal.
- 10 • Without any discretization of said received CPM signal along the sampling period, this signal is multiplied by $\cos(\omega_c T)$ and $\sin(\omega_c T)$ to yield an inphase component, $y^i(t) = \cos(\omega_c t) y(t)$ and a quadrature phase component, $y^q(t) = \sin(\omega_c t) y(t)$ respectively,
- For each successive time interval $nT \leq t \leq (n+1)T$ (with $0 \leq \tau \leq 1$) during which a symbol a_n is transmitted, carry out the following operations :
- 15 ➤ Define respectively for $y^i(t)$, and $y^q(t)$, the signals $y_n^i(\tau) = y^i((n+\tau)T)$ and $y_n^q(\tau) = y^q((n+\tau)T)$,
- Define signals $p(\tau)$ and $q(\tau)$ such that the derivative of $y_n^i(\tau) = y^i((n+\tau)T)$ and/or $y_n^q(\tau) = y^q((n+\tau)T)$ with respect to τ satisfy
- 20 a differential equation :

$$(\text{diff}) : \frac{d}{d\tau} y_n^i(\tau) = \pi h a_n y_n^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau) ,$$

- Treat said differential equation (diff) in order to :

- ✓ obtain a linear equation of the form : (lineq) $\alpha \{ \pi h a_n \} = \beta$, where α and β are coefficients which depend on iterated integrals of $y_n^i(\tau)$ and $y_n^q(\tau)$,
- 25 ✓ and retrieve for each time interval $[nT; (n+1)T]$ where a CPM signal is received a corresponding symbol a_n .

Preferred, but non limiting aspects of such method are the following

:

- $f(t)=\delta(t)$ (identity),
- in order to take into account the impulse response of said communication channel no equalization is carried out,
- said differential equation (diff) is valid wether or not the impulse response of said communication channel is neglected,
- the treatment of said differential equation (diff) comprises the following general steps :
 - Said differential equation (diff) is initially expressed in the temporal domain, as

$$(\text{diff}) : \frac{d}{d\tau} y_n^i(\tau) = \pi h a_n y_n^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau) ,$$
 - this differential equation (diff) is expressed in the frequency domain, through its Laplace transform, in order to obtain a corresponding equation (Laplace) : $s \hat{y}_n^i(s) = \pi h a_n \hat{y}_n^q(s) + \hat{v}_0(s) + \frac{\kappa}{s} + y_n^i(0)$, where s is the Laplace variable, $\hat{y}_n^i(s)$ and $\hat{y}_n^q(s)$ are the Laplace transforms of $y_n^i(\tau)$ and $y_n^q(\tau)$ respectively,
 - said treatments are carried out in the frequency domain :
 - a first treatment is applied to this Laplace transform in order to eliminate the constants, thus generating a new equation (4),
 - both members of the resulting equation (4) are then divided by s^λ , which corresponds to an iterated integration (λ times) in the temporal domain,
 - the modified equation is expressed in the temporal domain (through an inverse Laplace transform), and an equation of the form (lineq) is obtained,
- for said division λ is selected in order to be large enough to ensure that the resulting members of the modified equation after division do not comprise any multiplication by said Laplace variable,

- in said differential equation (diff), $p(\tau)$ and $q(\tau)$ are such that the signal $v(\tau) = \cos(\omega_c \tau)p(\tau) + \sin(\omega_c \tau)q(\tau)$, which incorporates the noise $b_n(\tau)$, is a high frequency component,
- as the iterated integrals of the high frequency component $v(\tau)$ tends to zero, this component, and consequently the original noise, has insignificant effect in the demodulation performance,
- in order to carry out said general steps, the following operations are performed :
 - The high frequency component $v(\tau)$, which includes the original noise is expressed as $v(\tau) = v_0(\tau) + \kappa$, where κ is a constant representing the mean value of $v(\tau)$, and $v_0(\tau)$ is a zero mean high frequency signal considered as a noise perturbation.
 - The differential equation (diff) is expressed in the operational domain, using the Laplace transform to obtain:

$$(4) : (\text{Laplace}) \quad s \hat{y}_n^i(s) = \pi h a_n \hat{y}_n^q(s) + \hat{v}_0(s) + \frac{\kappa}{s} + y_n^i(0),$$

where s is the Laplace variable, $\hat{y}_n^i(s)$ and $\hat{y}_n^q(s)$ are the Laplace transforms of $y_n^i(\tau)$ and $y_n^q(\tau)$ respectively.

- Both members of equation (Laplace) are multiplied by s^m , where $m \geq 1$ is an integer,
- The obtained equation is derived ℓ times with respect to said Laplace variable (s), with $\ell \geq m+1$, in order to eliminate the unknown constant κ and the initial condition $y_n^i(0)$, and generate a new equation of the form (equation (4))

$$(4) : \sum_{k=0}^{m+1} \alpha_k s^{m+1-k} \frac{d^{\ell-k} \hat{y}_n^i(s)}{ds^{\ell-k}} = \pi h a_n \sum_{k=0}^m \beta_k s^{m-k} \frac{d^{\ell-k} \hat{y}_n^q(s)}{ds^{\ell-k}} + \frac{d^\ell}{ds^\ell} \{s^m \hat{v}_0(s)\}$$

where

$$\alpha_k = \frac{\ell!}{k!(\ell-k)!} \frac{(m+1)!}{(m+1-k)!} \quad \text{and} \quad \beta_k = \frac{\ell!}{k!(\ell-k)!} \frac{(m)!}{(m-k)!},$$

- said new equation (4) is then treated as follows :

- (step i) both members of said new equation (4) are divided by s^λ where λ is an integer greater than $(m+2)$, thereby generating a modified equation,
- (step ii) Said modified equation is expressed in the temporal domain (through an inverse Laplace transform), by using the fact that:

✓ $\frac{d^k \hat{x}(s)}{ds^k}$ corresponds in the temporal domain to $(-1)^k \tau^k x(\tau)$,

and

- ✓ the division by s , corresponds in the temporal domain to an

integration – i.e. $s^{-\lambda} \frac{d^k \hat{x}(s)}{ds^k}$ can be expressed in the temporal domain as the integral, iterated at the order λ , of $(-1)^k \tau^k x(\tau)$,

- The coefficients α and β in equation (lineq) are obtained by applying (step i) and (step ii) to both members of said new equation (4).

Other aspects of the invention shall appear when reading the following description of the main steps carried out in the method according to the invention, made in reference to the drawings in which figure 2 is a graph which displays the bit error rate versus the signal to noise ratio per bit obtained by the demodulation method according to the invention (lower curve) in comparison with the theoretical bound for the optimum maximum likelihood demodulator (upper curve).

For a reason of clarity, the perturbation effect of the impulse response of the channel shall first be ignored in the following presentation : the convolution operated by the channel shall thus be set to identity i.e. $f(t)$ is set to the neutral element of the convolution operator, which is the delta Dirac impulse : $f(t) = \delta(t)$.

It will however be appreciated that the demodulation method of the invention is insensitive to the convolution effect of the channel.

This method is a demodulation method for demodulating a received CPM signal ($y(t)$) having a sampling period T , a modulation index h and a carrier frequency (ω_c) and received from a communication channel, said received CPM signal being, in each time interval $[nT, (n+1)T]$ and for
 5 $f(t)=\delta(t)$, of the general form :

$$y(t)=A\cos(\{\omega_c T+\pi h a_n\}t+\varphi_n)+b(t)$$

where $t = \text{time} = (n + \tau)T, 0 \leq \tau \leq 1$ and where $\varphi_n = \pi h \sum_{k=0}^{n-1} a_k - n\omega_c T$.

Note that this general form of $y(t)$ can also be expressed as :

$$y(t)=A \int_{-\infty}^{\infty} f(\lambda) \cos \left(\omega_c (t-\lambda) + \pi h a_n \left[\frac{t-\lambda}{T} - n \right] + \pi h \sum_{k=-\infty}^{n-1} a_k \right) d\lambda + b(t),$$

10 this last expression being more general as it explicitly comprises $f(t)$ (i.e. it does not assume that $f(t)=\delta(t)$).

As mentioned in the introduction of this text with reference to figure 1 (which general scheme is applicable also for the invention), the received CPM signal results from the addition of an original modulated signal $u(t)$ and
 15 a noise component $b(t)$.

The original modulated signal $u(t)$ has been obtained by a CPM modulation – e.g. a Continuous Phase Frequency Shift Keying (CPFSK) method.

For $nT \leq t \leq (n+1)T$, $u(t)$ can thus be expressed as (equation (1)) :

$$20 \quad (1) \quad u(t) = u((n+\tau)T) = u_n(\tau) = A \cos(\omega_n \tau + \varphi_n); \quad \tau = t/T - n,$$

with $\omega_n = \omega_c T + \pi h a_n$

and likewise, the received $y(t)$ can be expressed, within the same time interval, as

$$y(t) = y((n+\tau)T) = y_n(\tau) = A \cos(\omega_n \tau + \varphi_n) + b_n(\tau); \quad \tau = t/T - n$$

25 where $b_n(\tau) = b(t = (n+\tau)T)$ represents the noise disturbance in the time interval $[nT, (n+1)T]$

The sequence $\{a_n\}$ forms a series of symbols which are themselves generated from an initial signal $s(t)$ to be transmitted via the communication

channel 30. Here again, the communication channel is any type of known communication channel, e.g. a wireless channel such as a radio channel.

The noise component ($b(t)$) is generated by the transmission by said communication channel.

5 And the "received CPM signal" received at the receiver associated to the demodulator 40 is of the form $y(t) = u(t) + b(t)$.

For carrying out the method of the invention, the following steps are performed in real time as said received CPM signal is received at the receiver :

- 10 • Said received CPM signal $y(t)$ is received as a continuous signal.
- Without any discretization of $y(t)$ along the sampling period , this signal is multiplied by $\cos(\omega_c T)$ and $\sin(\omega_c T)$ to yield an inphase component, $y^i(t)$, and a quadrature phase component, $y^q(t)$, respectively. These components are therefore defined as $y^i(t) = \cos(\omega_c t)y(t)$ and
- 15 $y^q(t) = \sin(\omega_c t)y(t)$,
- More precisely, these components $y^i(t)$, and $y^q(t)$, are defined for each time interval $nT \leq t \leq (n+1)T$, corresponding to the time interval $[nT, (n+1)T]$, during which the symbol a_n is transmitted :
- For that purpose $t = (n + \tau)T$; $0 \leq \tau \leq 1$, is set, and
- 20 ➤ $y_n^i(\tau) = y^i((n + \tau)T)$ and $y_n^q(\tau) = y^q((n + \tau)T)$ are defined,
- the following operations shall then be carried out within each successive time interval $nT \leq t \leq (n+1)T$:
- The derivative of the signal $y_n^i(\tau) = y^i((n + \tau)T)$ with respect to τ satisfies the following relation :

25 (diff)
$$\frac{d}{d\tau} y_n^i(\tau) = \pi h a_n y_n^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau)$$

where $p(\tau)$ and $q(\tau)$ are such that the signal $v(\tau) = \cos(\omega_c \tau)p(\tau) + \sin(\omega_c \tau)q(\tau)$, which incorporates the noise $b_n(\tau)$, is a high frequency component.

- It is to be noted that this differential equation (diff) is valid whether or not the channel's impulse response $f(t)$ is neglected (i.e. $f(t)=\delta(t)$) – this shall be exposed in a separate section at the end of the present description,
- 5 ➤ Furthermore, the equalization step carried out with some known demodulation methods is needless with the invention.

- This differential equation (diff) is then treated in order to :
 - obtain a linear equation of the form :

$$(\text{lineq}) \quad \alpha \{ \pi h a_n \} = \beta ,$$

- 10 where α and β are coefficients which depend on iterated integrals of $y_h^i(\tau)$ and $y_h^q(\tau)$. As the iterated integrals of the high frequency component $v(\tau)$ tends to zero, this component, and consequently the original noise, has insignificant effect in the demodulation performance.

- 15 ➤ and retrieve for each time (t) interval $[nT; (n+1)T]$ where a CPM signal is received a corresponding symbol a_n .

It is specified that an equivalent differential equation can be exploited for $y_n^q(\tau) = y^q((n+\tau)T)$.

In such equation :

- 20 • $y^i(t)$, would be replaced by $y^q(t)$, and vice-versa,
- the first member of the right-handside of the equation would be multiplied by a negative sign (-),
 - the signals $p(\tau)$ and $q(\tau)$ would be different.

And in order to carry out the demodulation, it is possible to treat
25 either of these two differential equations (or both in parallel).

The treatment of equation (diff) comprises the following general steps :

- equation (diff) is initially expressed in the temporal domain, as

$$\frac{d}{d\tau} y_n^i(\tau) = \pi h a_n y_n^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau) ,$$

- this equation (diff) is expressed in the frequency domain, through its Laplace transform, in order to obtain a corresponding equation (noted (Laplace) below),
- the following treatments are then carried out in this frequency domain :
 - a first treatment is applied to this Laplace transform in order to eliminate the constants (represented as will be explained below by κ and by $y_n^i(0)$). A new equation (noted (4) below) is thus obtained,
 - both members of the resulting equation are then divided by s^λ , which corresponds to an iterated integration (λ times) in the temporal domain. For this division, λ is selected in order to be large enough to ensure that the resulting members of the modified equation after division do not comprise any multiplication by s ,
- the modified equation is expressed in the temporal domain (through an inverse Laplace transform), and an equation of the form (lineq) is obtained.

More precisely, in order to carry out the above steps, a preferred – but non limitative – approach is as follows :

- The high frequency component $v(\tau)$, which includes the original noise is expressed as $v(\tau) = v_0(\tau) + \kappa$, where κ is a constant representing the mean value of $v(\tau)$, and $v_0(\tau)$ is a zero mean high frequency signal considered as a noise perturbation.
- The differential equation (diff) is expressed in the operational domain, using the Laplace transform to obtain:

$$\text{(Laplace)} \quad s\hat{y}_n^i(s) = \pi h a_n \hat{y}_n^q(s) + \hat{v}_0(s) + \frac{\kappa}{s} + y_n^i(0)$$

- where s is the Laplace variable, $\hat{y}_n^i(s)$ and $\hat{y}_n^q(s)$ are the Laplace transforms of $y_n^i(\tau)$ and $y_n^q(\tau)$ respectively.
- Both members of equation (Laplace) are multiplied by s^m , where $m \geq 1$ is an integer,

- The obtained equation is derived ℓ times with respect to s , with $\ell \geq m+1$, in order to eliminate the unknown constant κ and the initial condition $y_n^i(0)$. Therefore is obtained an equation of the form (equation (4))

$$(4) \quad \sum_{k=0}^{m+1} \alpha_k s^{m+1-k} \frac{d^{\ell-k} \hat{y}_n^i(s)}{ds^{\ell-k}} = \pi h a_n \sum_{k=0}^m \beta_k s^{m-k} \frac{d^{\ell-k} \hat{y}_n^q(s)}{ds^{\ell-k}} + \frac{d^\ell}{ds^\ell} \{s^m \hat{v}_0(s)\}$$

5 where $\alpha_k = \frac{\ell!}{k!(\ell-k)!} \frac{(m+1)!}{(m+1-k)!}$ and $\beta_k = \frac{\ell!}{k!(\ell-k)!} \frac{(m)!}{(m-k)!}$

- This modified equation (4) is then treated as follows :

➤ (step i) in order to eliminate from this equation (4) the terms s^{m+1-k} and s^{m-k} , both members of equation (4) are divided by s^λ where λ is an integer greater than $(m+2)$. This operation (which corresponds to an iterated integration in the temporal domain) generates a modified equation,

➤ (step ii) Said modified equation is expressed in the temporal domain (through an inverse Laplace transform), by using the fact that – for a generic function $x(\tau)$ having a Laplace transform expressed as $\hat{x}(s)$:

15 ✓ $\frac{d^k \hat{x}(s)}{ds^k}$ corresponds in the temporal domain to $(-1)^k \tau^k x(\tau)$, and

✓ the division by s , corresponds in the temporal domain to an integration – i.e. $s^{-\lambda} \frac{d^k \hat{x}(s)}{ds^k}$ can be expressed in the temporal

domain as the integral, iterated at the order λ , of $(-1)^k \tau^k x(\tau)$,

- The coefficients α and β in equation (lineq) are obtained by applying (step i) and (step ii) to both members of equation (4). Note that the noise term, $\frac{d^\ell}{ds^\ell} \{s^m \hat{v}_0(s)\}$, in the right hand side of equation (4) is annihilated by the application of these steps as it is a high-frequency term.

Numerical example

25

Let us take a concrete example to fix the notations. Let $m=1$, $\ell=2$ and $\lambda=3$. Equation (4) then becomes :

$$s^2 \frac{d^2 \hat{y}_n^i(s)}{ds^2} + 4s \frac{d\hat{y}_n^i(s)}{ds} + 2\hat{y}_n^i(s) = \pi h a_n \left(s \frac{d^2 \hat{y}_n^q(s)}{ds^2} + 2 \frac{d\hat{y}_n^q(s)}{ds} \right) + s \frac{d^2 \hat{v}_0(s)}{ds^2} + 2 \frac{d\hat{v}_0(s)}{ds}$$

Dividing both sides of this equation by s^3 yields

(4b)

5 :

$$\frac{1}{s} \frac{d^2 \hat{y}_n^i(s)}{ds^2} + \frac{4}{s^2} \frac{d\hat{y}_n^i(s)}{ds} + \frac{2}{s^3} \hat{y}_n^i(s) = \pi h a_n \left(\frac{1}{s^2} \frac{d^2 \hat{y}_n^q(s)}{ds^2} + \frac{2}{s^3} \frac{d\hat{y}_n^q(s)}{ds} \right) + \frac{1}{s^2} \frac{d^2 \hat{v}_0(s)}{ds^2} + \frac{2}{s^3} \frac{d\hat{v}_0(s)}{ds}$$

This equation is now expressed back in the time domain by noting

that the time domain equivalent of a term having a form $\frac{1}{s^k} \frac{d^j \hat{x}(s)}{ds^j}$ is given

10 by

$$\int_0^t dt_1 \int_0^{t_1} dt_2 \dots \int_0^{t_{k-1}} (-1)^j \tau^j x(\tau) d\tau = \frac{(-1)^j}{(k-1)!} \int_0^t (t-\tau)^{k-1} \tau^j x(\tau) d\tau$$

Equation (4b) then reads in the time domain as:

(lin) :

$$\int_0^\eta (6\tau^2 - 6\eta\tau + \eta^2) y_n^i(\tau) d\tau = \left(\int_0^\eta \tau(\eta-\tau)(2\tau-\eta) y_n^q(\tau) d\tau \right) \pi h a_n + \int_0^\eta \tau(\eta-\tau)(2\tau-\eta) v_0(\tau) d\tau$$

15

where the integration time η is lower than 1, $0 < \eta \leq 1$, since the signals $y_n^i(\tau)$ and $y_n^q(\tau)$ are defined for $\tau \in [0,1]$.

This integration time η may be chosen very small ; this explain why the demodulation method may be implemented on-line.

20

The last term of the right hand side of this equation corresponds to iterated integrals of the high frequency component $v_0(\tau)$. Therefore, it has an insignificant effect on equation (lin). Taking this fact into account, one finally obtains the linear equation (lineq) with

$$\alpha = \int_0^\eta \tau(\eta-\tau)(2\tau-\eta) y_n^q(\tau) d\tau$$

25

and

$$\beta = \int_0^{\eta} (6\tau^2 - 6\eta\tau + \eta^2) y_n^i(\tau) d\tau$$

The n^{th} symbol a_n is thus recovered by the explicit formula :

$$(\text{demod}) : a_n = \frac{\int_0^{\eta} (6\tau^2 - 6\eta\tau + \eta^2) y_n^i(\tau) d\tau}{\pi h \int_0^{\eta} \tau(\eta - \tau)(2\tau - \eta) y_n^q(\tau) d\tau}$$

Note that it is only for the purpose of computing numerically the
5 integrals appearing in the above equation that the sampling of the received signal $y(t)$ is necessary.

Simulation example

10 A simulation example for the demodulation of a CPFSK modulated signal, received over an additive white Gaussian noise channel shall now be presented with the following parameters :

$$\frac{\omega_c}{2\pi} = 900 \text{ MHz}; \quad T = 1/8000 \text{ s}; \quad h = 0.725 \quad \text{and} \quad \text{the bearing}$$

information symbols $\{a_n\}$ are drawn from an M-ary alphabet with $M=8$.

15 Figure 2 displays the bit error rate versus the signal to noise ratio per bit, obtained by the demodulation method according to the invention (lower curve) in comparison with the theoretical bound for the optimum maximum likelihood demodulator (upper curve). In the presence of a high level of noise, the demodulator given in formula (demod) is particularly
20 performant.

Case where the condition $f(t)=\delta(t)$ is not met

This section shows that equation (diff) also holds for any channel
25 having an impulse response $f(t)$ ($f(t)$ being possibly different from identity).

This corresponds to a more realistic situation where the distortion induced by the communication medium (i.e. the channel) is more complicated than a simple additive noise.

The received signal then reads as :

$$(5) \quad (\text{output}) \quad y(t) = f(t) * u(t) + b(t) = \int_{-\infty}^{\infty} f(\tau) u(t-\tau) d\tau + b(t),$$

where $f(t)$ still denotes the channel's impulse response, $u(t)$ is the input signal and $b(t)$ an additive noise.

To proceed, let us recall that the general form of the modulated signal $u(t)$ is given by:

$$(10) \quad (\text{input}) \quad u(t) = A \Re \left\{ e^{i\omega_c t} e^{i2\pi h \sum_{k \in \mathbb{Z}} a_k q(t-kT)} \right\}$$

where $\Re\{z\}$ stands for the real part of the complex number z , and A is a constant amplitude.

For a CPFSK modulation, the waveform $q(t)$ is given by

$$q(t) = \begin{cases} 0 & ; \quad t < 0, \\ 1/2 T & ; \quad 0 \leq t \leq T \\ 1/2 & \quad t \geq T \end{cases}$$

(15) Thus, for $nT \leq t \leq (n+1)T$, the input signal $u(t)$ becomes

$$u(t) = A \cos \left(\omega_c t + \pi h a_n \left[\frac{t}{T} - n \right] + \pi h \sum_{k=-\infty}^{n-1} a_k \right), \text{ which corresponds to}$$

equation (1).

The expression of the received signal $y(t)$ in equation (output) is thus

(20) (outputb)

$$y(t) = A \int_{-\infty}^{\infty} f(\lambda) \cos \left(\omega_c (t-\lambda) + \pi h a_n \left[\frac{t-\lambda}{T} - n \right] + \pi h \sum_{k=-\infty}^{n-1} a_k \right) d\lambda + b(t)$$

In order to simplify the expressions to follow, we set

$$(\text{def}) \quad g_n(t, \lambda) = \pi h a_n \left[\frac{t-\lambda}{T} - n \right] + \pi h \sum_{k=-\infty}^{n-1} a_k$$

Now, one can derive the explicit form of the inphase component as:

(25)

$$y^i(t) = \cos(\omega_c t) y(t) = \frac{A}{2} \int_{-\infty}^{\infty} f(\lambda) \cos(\omega_c \lambda - g_n(t, \lambda)) d\lambda + \frac{A}{2} c(t) + \cos(\omega_c t) b(t)$$

where $c(t)$ is defined by

$$c(t) = \cos(2\omega_c t) \int_{-\infty}^{\infty} f(\lambda) \cos(\omega_c \lambda - g_n(t, \lambda)) d\lambda + \sin(2\omega_c t) \int_{-\infty}^{\infty} f(\lambda) \sin(\omega_c \lambda - g_n(t, \lambda)) d\lambda$$

Likewise, the quadrature phase component reads as:

5

$$y^q(t) = \sin(\omega_c t) y(t) = \frac{A}{2} \int f(\lambda) \sin(\omega_c \lambda - g_n(t, \lambda)) d\lambda + \frac{A}{2} e(t) + \sin(\omega_c t) b(t)$$

where :

$$e(t) = \sin(2\omega_c t) \int f(\lambda) \cos(\omega_c \lambda - g_n(t, \lambda)) d\lambda - \cos(2\omega_c t) \int f(\lambda) \sin(\omega_c \lambda - g_n(t, \lambda)) d\lambda$$

Taking the derivative of $y^i(t)$ with respect to t , allows one to obtain

10

(diffb) :

$$\frac{dy^i(t)}{dt} = \frac{dg_n(t, \lambda)}{dt} y^q(t) - \frac{dg_n(t, \lambda)}{dt} \left\{ \frac{A}{2} e(t) + \sin(\omega_c t) b(t) \right\} + \frac{d}{dt} \left\{ \frac{A}{2} c(t) + \cos(\omega_c t) b(t) \right\}$$

Now, observe that the signals $\frac{A}{2} c(t) + \cos(\omega_c t) b(t)$ and

$\frac{A}{2} e(t) + \sin(\omega_c t) b(t)$ (which incorporate the noise) are high frequency components.

15

Hence for their contributions in the above differential equation (diffb) they are considered as high frequency noise perturbation which may be put in the form $\cos(\omega_c t) \tilde{p}(t) + \sin(\omega_c t) \tilde{q}(t)$ for some (noisy) signals $\tilde{p}(t)$ and $\tilde{q}(t)$.

20

Finally, using the definition (def) of $g_n(t, \lambda)$, the differential equation (diffb) becomes

$$\frac{dy^i(t)}{dt} = \frac{\pi h a_n}{T} y^q(t) + \cos(\omega_c t) \tilde{p}(t) + \sin(\omega_c t) \tilde{q}(t)$$

The differential equation (diff) now follows by using the change of variables:

$$\tau = \frac{t}{T} - n, \quad y_n^i(\tau) = y^i((n + \tau)T) = y^i(t) \quad \text{and} \quad y_n^q(\tau) = y^q((n + \tau)T) = y^q(t) :$$

25

$$(diff) \quad \frac{dy_n^i(\tau)}{d\tau} = \pi h a_n y^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau)$$

Additional comments

The method described above is carried out by the demodulator 40
 5 for each time interval defined by nT and $(n+1)T$. This method allows one to determine the symbols a_n on a continuous symbol-per-symbol basis, in real time if desired, as the signal $y(t)$ is received.

This method is based on a deterministic approach (i.e. it is an explicit method which is not based on a probabilistic approach).

10 Furthermore, it is to be noted that no discretization of the received CPM signal is carried out before actually demodulating said received CPM signal. Indeed, the method according to the invention does not require such a priori discretization and allows working on the "crude" and continuous received CPM signal $y(t)$.

15 Moreover, in the case of the invention no discretization at all is required along the sampling period – contrary to what is done in the known methods.

In addition, the following advantages have to be noted with respect to the invention.

20 The demodulation method of this invention is valid even if the communication channel introduces an intersymbol interference, in addition to the additive noise perturbation, and this without requiring any equalization step..

25 The performance of the demodulation presented in this invention are not affected by a carrier phase shift (non coherent detection) as opposed to the existing methods.

By their construction, each of the parameters $y_n^i(\tau) = y^i((n+\tau)T)$ and $y_n^q(\tau) = y^q((n+\tau)T)$ respectively comprise a high-frequency component and a low-frequency component.

30 And the successive integrations carried out on these parameters allow eliminating the high frequency components of these parameters (i.e. of the received CPM signal) – therefore acting as a lowpass filter.

These high-frequency components typically contain the noise component $b(t)$, as well as other noise components already present in the signal $u(t)$. And the successive integrations carried out thus allow one to eliminate such noise components.

5 A low-pass filter is therefore not needed – even if it can be added if desired.

It is also to be noted that contrary to the known methods where the arithmetic complexity grows exponentially with the size of the register used for elaborating the series $\{a_n\}$, such register size has no influence on the
10 method according to the invention and such method is therefore particularly well-suited for demodulating with very high rates.

Since the method according to the invention is carried out on a symbol-per-symbol basis, the determination of a symbol a_n is not influenced by the earlier determination of previous symbols – and there is therefore no
15 risk of error accumulation.

Since the symbols a_n are determined by an explicit method, without requiring any optimization, the method according to the invention can be carried out in real-time. The time interval required for demodulating a given symbol a_n is smaller than the period T of the series of the symbols
20 themselves.

Finally, the features of the method according to the invention make it well adapted for demodulating in a coherent manner, as well as in a non-coherent manner.

5 CLAIMS

1. Method for demodulating a received CPM signal ($y(t)$) having a sampling period T and a carrier frequency (ω_c) and received from a communication channel, said received CPM signal being of the general form :

$$y(t) = A \int_{-\infty}^{\infty} f(\lambda) \cos \left(\omega_c(t-\lambda) + \pi h a_n \left[\frac{t-\lambda}{T} - n \right] + \pi h \sum_{k=-\infty}^{n-1} a_k \right) d\lambda + b(t)$$

where the convolution by the impulse response of the channel $f(t)$ represents the perturbation effect of the channel, and where $b(t)$ is an additive noise; for the n th sampling interval of the signal :

- 15 $t = \text{time} = (n + \tau)T$, $0 \leq \tau \leq 1$, h : modulation index, and $\varphi_n = \pi h \sum_{k=0}^{n-1} a_k - n\omega_c T$, said received CPM signal ($y(t)$) being obtained from :

➤ an original modulated signal ($u(t)$) generated from a series of symbols (a_n), said series being itself generated from an initial signal to be transmitted via said communication channel, said original modulated

- 20 signal being of the general form for $nT \leq t \leq (n+1)T$:

(equation(1)). $u(t) = A \cos(\omega_n \tau + \varphi_n)$ $\tau = t/T - n$

with : $\omega_n = \omega_c T + \pi h a_n$, and

➤ a signal ($b(t)$) corresponding to a noise generated by the transmission by said communication channel,

- 25 said method being **characterized in that** it comprises the following steps carried out in real time as said received CPM signal is received :

- Said received CPM signal ($y(t)$) is received as a continuous signal.

- Without any discretization of said received CPM signal along the sampling period, this signal is multiplied by $\cos(\omega_c T)$ and $\sin(\omega_c T)$ to yield an inphase component, $y^i(t) = \cos(\omega_c t) y(t)$ and a quadrature phase component, $y^q(t) = \sin(\omega_c t) y(t)$ respectively,
- 5 • For each successive time interval $nT \leq t \leq (n+1)T$ (with $0 \leq \tau \leq 1$) during which a symbol a_n is transmitted, carry out the following operations :
- Define respectively for $y^i(t)$, and $y^q(t)$, the signals $y_n^i(\tau) = y^i((n+\tau)T)$ and $y_n^q(\tau) = y^q((n+\tau)T)$,
 - Define signals $p(\tau)$ and $q(\tau)$ such that the derivative of $y_n^i(\tau) = y^i((n+\tau)T)$ and/or $y_n^q(\tau) = y^q((n+\tau)T)$ with respect to τ satisfy a differential equation :
- 10 (diff) : $\frac{d}{d\tau} y_n^i(\tau) = \pi h a_n y_n^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau)$,
- Treat said differential equation (diff) in order to :
 - ✓ obtain a linear equation of the form : (lineq) $\alpha \{ \pi h a_n \} = \beta$, where
- 15 α and β are coefficients which depend on iterated integrals of $y_n^i(\tau)$ and $y_n^q(\tau)$,
- ✓ and retrieve for each time interval $[nT; (n+1)T]$ where a CPM signal is received a corresponding symbol a_n .
- 20 2. Method according to claim 1 characterized in that $f(t) = \delta(t)$.
3. Method according to any preceding claim characterized in that in order to take into account the impulse response ($f(t)$) of said communication channel no equalization is carried out.
- 25 4. Method according to any preceding claim characterized in that said differential equation (diff) is valid whether or not the impulse response ($f(t)$) of said communication channel is neglected.

5. Method according to any preceding claim characterized in that the treatment of said differential equation (diff) comprises the following general steps :

- Said differential equation (diff) is initially expressed in the temporal domain, as

$$(\text{diff}) : \frac{d}{d\tau} y_n^i(\tau) = \pi h a_n y_n^q(\tau) + \cos(\omega_c \tau) p(\tau) + \sin(\omega_c \tau) q(\tau),$$

- this differential equation (diff) is expressed in the frequency domain, through its Laplace transform, in order to obtain a corresponding equation (Laplace) : $s \hat{y}_n^i(s) = \pi h a_n \hat{y}_n^q(s) + \hat{v}_0(s) + \frac{\kappa}{s} + y_n^i(0)$, where s is

the Laplace variable, $\hat{y}_n^i(s)$ and $\hat{y}_n^q(s)$ are the Laplace transforms of $y_n^i(\tau)$ and $y_n^q(\tau)$ respectively,

- said treatments are carried out in the frequency domain :
 - a first treatment is applied to this Laplace transform in order to eliminate the constants, thus generating a new equation (4),
 - both members of the resulting equation (4) are then divided by s^λ , which corresponds to an iterated integration (λ times) in the temporal domain,
- the modified equation is expressed in the temporal domain (through an inverse Laplace transform), and an equation of the form (lineq) is obtained.

6. Method according to the preceding claim characterized in that for said division λ is selected in order to be large enough to ensure that the resulting members of the modified equation after division do not comprise any multiplication by said Laplace variable (s).

7. Method according to any of the two preceding claims characterized in that in said differential equation (diff), $p(\tau)$ and $q(\tau)$ are such that the

signal $v(\tau) = \cos(\omega_c \tau)p(\tau) + \sin(\omega_c \tau)q(\tau)$, which incorporates the noise $b_n(\tau)$, is a high frequency component.

8. Method according to the preceding claim characterized in that as the iterated integrals of the high frequency component $v(\tau)$ tends to zero, this component, and consequently the original noise, has insignificant effect in the demodulation performance.

9. Method according to any of the four preceding claims characterized in that in order to carry out said general steps, the following operations are performed :

- The high frequency component $v(\tau)$, which includes the original noise is expressed as $v(\tau) = v_0(\tau) + \kappa$, where κ is a constant representing the mean value of $v(\tau)$, and $v_0(\tau)$ is a zero mean high frequency signal considered as a noise perturbation.
- The differential equation (diff) is expressed in the operational domain, using the Laplace transform to obtain:

$$(\text{Laplace}) \quad s\hat{y}_n^i(s) = \pi h a_n \hat{y}_n^q(s) + \hat{v}_0(s) + \frac{\kappa}{s} + y_n^i(0),$$

where s is the Laplace variable, $\hat{y}_n^i(s)$ and $\hat{y}_n^q(s)$ are the Laplace transforms of $y_n^i(\tau)$ and $y_n^q(\tau)$ respectively.

- Both members of equation (Laplace) are multiplied by s^m , where $m \geq 1$ is an integer,
- The obtained equation is derived ℓ times with respect to said Laplace variable (s), with $\ell \geq m+1$, in order to eliminate the unknown constant κ and the initial condition $y_n^i(0)$, and generate a new equation of the form (equation (4))

$$(4) : \sum_{k=0}^{m+1} \alpha_k s^{m+1-k} \frac{d^{\ell-k} \hat{y}_n^i(s)}{ds^{\ell-k}} = \pi h a_n \sum_{k=0}^m \beta_k s^{m-k} \frac{d^{\ell-k} \hat{y}_n^q(s)}{ds^{\ell-k}} + \frac{d^\ell}{ds^\ell} \{s^m \hat{v}_0(s)\}$$

where

$$\alpha_k = \frac{\ell!}{k!(\ell-k)!} \frac{(m+1)!}{(m+1-k)!} \quad \text{and} \quad \beta_k = \frac{\ell!}{k!(\ell-k)!} \frac{(m)!}{(m-k)!},$$

- said new equation (4) is then treated as follows :
 - (step i) both members of said new equation (4) are divided by s^λ where λ is an integer greater than $(m+2)$, thereby generating a modified equation,
 - 5 ➤ (step ii) Said modified equation is expressed in the temporal domain (through an inverse Laplace transform), by using the fact that:
 - ✓ $\frac{d^k \hat{x}(s)}{ds^k}$ corresponds in the temporal domain to $(-1)^k \tau^k x(\tau)$,
 - and
 - 10 ✓ the division by s , corresponds in the temporal domain to an integration – i.e. $s^{-\lambda} \frac{d^k \hat{x}(s)}{ds^k}$ can be expressed in the temporal domain as the integral, iterated at the order λ , of $(-1)^k \tau^k x(\tau)$,
- The coefficients α and β in equation (lineq) are obtained by
 - 15 applying (step i) and (step ii) to both members of said new equation (4).

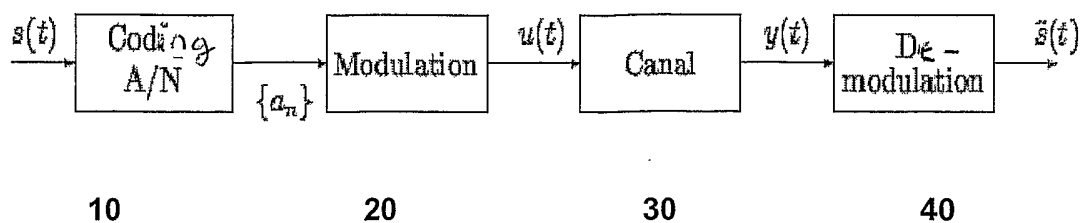


FIG. 1

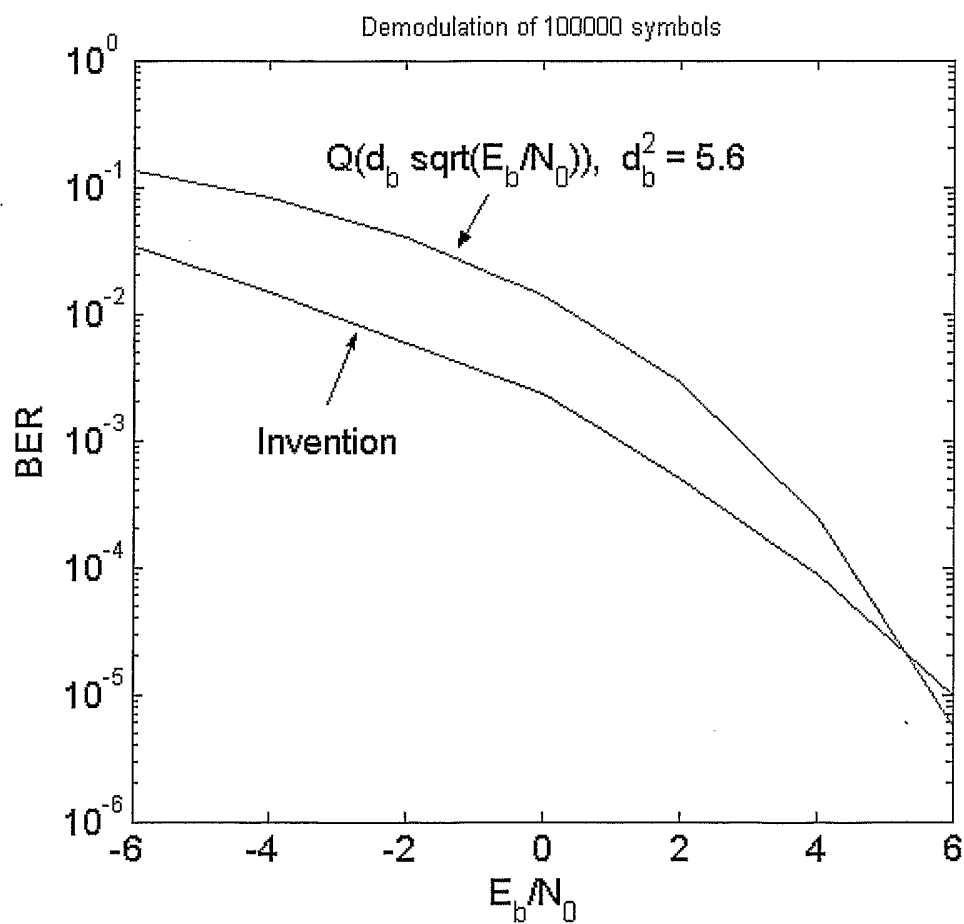


FIG. 2

INTERNATIONAL SEARCH REPORT

International Application No
PCT/IB2004/004416

A. CLASSIFICATION OF SUBJECT MATTER

IPC 7 H04L27/22

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	FLIESS M ET AL: "A new computational approach for the demodulation of CPM signals" SIGNAL PROCESSING AND INFORMATION TECHNOLOGY, 2003. ISSPIT 2003. PROCEEDINGS OF THE 3RD IEEE INTERNATIONAL SYMPOSIUM ON DARMSTADT, GERMANY 14-17 DEC. 2003, PISCATAWAY, NJ, USA, IEEE, 14 December 2003 (2003-12-14), pages 391-394, XP010729174 ISBN: 0-7803-8292-7 abstract page 392 ----- -/--	1-9

☒ Further documents are listed in the continuation of box C.

☒ Patent family members are listed in annex.

° Special categories of cited documents :

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Date of the actual completion of the international search

29 July 2005

Date of mailing of the international search report

08/08/2005

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INTERNATIONAL SEARCH REPORT

International Application No
PCT/IB2004/004416

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

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X	HU V ED - INSTITUTE OF ELECTRICAL AND ELECTRONICS ENGINEERS: "AN OPTIMIZED DIGITAL MODEM ARCHITECTURE FOR VLSI IMPLEMENTATION" COMMUNICATIONS : CONNECTING THE FUTURE. SAN DIEGO, DEC. 2 - 5, 1990, PROCEEDINGS OF THE GLOBAL TELECOMMUNICATIONS CONFERENCE AND EXHIBITION(GLOBECOM), NEW YORK, IEEE, US, vol. VOL. 3, 2 December 1990 (1990-12-02), pages 1691-1695, XP000218860 ISBN: 0-87942-632-2 page 1691, left-hand column, last paragraph - right-hand column, paragraph 3 page 1692, left-hand column, last paragraph - page 1694, left-hand column, paragraph 3; figures 3-5 equations (1),(2),(6),(15),(18)	1
A	AULIN T ET AL: "CONTINUOUS PHASE MODULATION - PART II: PARTIAL RESPONSE SIGNALING" IEEE TRANSACTIONS ON COMMUNICATIONS, IEEE INC. NEW YORK, US, vol. COM-29, no. 3, March 1981 (1981-03), pages 210-225, XP000758696 ISSN: 0090-6778 cited in the application pages 221-223	1-9

INTERNATIONAL SEARCH REPORT

Information on patent family members

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