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DIGITAL MODULATION PROCESS AND MODULATOR IMPLEMENTING THE PROCESS

This invention concerns digital radio communication systems. It concerns, in particular, the methods of modulation implemented in such systems.

Digital modulation is generally designed to combine the following three requirements: a high transmission rate, minimum spectrum occupancy and a low transmission error rate under various operating conditions.

Various methods were proposed in the past with a view to achieving a high transmission rate on a channel with reduced spectral bandwidth (transmission rate exceeding 1 bit/s/Hz).

The first group of methods uses multi-level frequency modulation as a basis, together with adequate filtering of the modulating signal (e.g. Gaussian filtering used with GMSK modulation) in order to reduce adjacent channel interference. These methods have the advantage that they are easily applied, and result in modulated signals of constant envelope. They consequently permit transmitters to be fitted with power amplifiers which operate in the saturated state. These amplifiers are readily available, cheap and very efficient. However, in order to comply with constraints relating to adjacent channel interference, the modulation index must be limited considerably, and the modulating signal thoroughly filtered. This causes the symbol spacing to be reduced, and this adversely affects the noise immunity of the modulation. In other words, the sensitivity of the radio receivers is limited.

Another group of methods uses phase-shift keying (PSK) and, if necessary, differential phase shift keying (DPSK) as a basis, and the resulting signal is filtered to ensure that standards relating to adjacent channel interference are complied with. In general, a filter satisfying the Nyquist criterion is used in order to limit inter-symbol interference. These methods generally provide satisfactory sensitivity at the expense of a large variation in the amplitude of the radio signal. Very linear amplifiers are therefore necessary, and they are difficult to design and set up. In addition, they are

generally inefficient, and this seriously affects the autonomy of mobile stations. A non-linear amplifier can be used in conjunction with a linearising method, but such method complicates a transmitter very considerably if there are large envelope variations.

5 Other solutions have also been proposed, e.g. in U.S. Patents 5,642,384 and 5,311,552, where an appropriate choice of a constellation and of a coded modulation process prevents transitions in the constellation for which the phase change is relatively large. This permits the variation in amplitude of a radio signal to be reduced to values compatible with the
10 characteristics of amplifiers which are easier to design. However, the reduction in amplitude is achieved at the expense of a considerable reduction in the symbol spacing, which is very difficult to compensate by coding gains, in particular in the error rate range of the greatest importance to speech communications, i.e. for bit error rates (BER) of the order of 10^{-2} , especially
15 when the channel is affected by fading (Rayleigh fading).

An object of the present invention is to propose a digital modulation group permitting joint optimisation of noise immunity, even in a channel affected by fading, adjacent channel interference, and variation in amplitude of the radio signal.

20 The invention thus proposes a digital modulation process wherein the successive symbols of a digital stream are converted into phase increments, an accumulated phase is obtained by adding the successive phase increments, a modulating phase is obtained by filtering the accumulated phase, a complex signal is produced whose argument represents the
25 modulating phase, two phase quadrature radio waveforms are respectively modulated on the basis of said complex signal, and a radio signal resulting from a combination of the two modulated waveforms is transmitted. According to the invention, said complex signal is digitally filtered, and digital signals obtained from the real and imaginary components of the digitally filtered
30 complex signal are converted into analog form before being respectively subjected to anti-aliasing analog filtering and then mixed with the two radio waveforms.

Said digital signals obtained from the real and imaginary



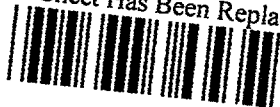


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components of the digitally filtered complex signal typically consist of the real and imaginary components themselves. However, if an amplifier linearising process is used, by pre-distortion for example (see European patent application No. 0 797 293), the real and imaginary components may be
5 subject to correction before being converted into analog form. The use of a linearising process is not included directly in this invention. In many cases, the invention will permit such a process to be dispensed with. In other cases, it will permit the use of such processes to be simplified considerably (for example, by not taking account of phase changes), in view of the small
10 variations in the signal envelope permitted by an appropriate choice of parameters for filtering the accumulated phase and said real and imaginary components. The criteria for this selection will be specified further on.

The invention permits digital radio communication systems, in particular professional radio communication systems, to be implemented in
15 accordance with applicable standards relating to adjacent channel interference, and provides unequalled sensitivity and thus radio range, using power amplifier components which are readily available on the market and have a high power efficiency.

Another aspect of the invention relates to a digital modulator,
20 including means for converting successive symbols of a digital stream into phase increments, a summator which accumulates the successive phase increments to produce an accumulated phase, a phase filter receiving the accumulated phase and producing a modulating phase, means for producing a complex signal whose argument represents the modulating phase, and a
25 modulator for respectively modulating two phase quadrature radio waveforms on the basis of said complex signal, and for transmitting a radio signal resulting from a combination of the two modulated waveforms, the modulator comprising a digital filter to which the complex signal is applied, analog-digital converters respectively processing the digital signals obtained from the real
30 and imaginary components of the digitally filtered complex signal, anti-aliasing analog filters receiving the output signals from the analog-digital converters, and two mixers each receiving one of the two radio waveforms and the output signal from one of the two anti-aliasing filters.



Other features and advantages of the invention will become apparent from the following description of non-limiting embodiments, with reference to the attached drawings where:

- Figures 1 and 2 are block diagrams of a digital modulator in accordance with the invention and an associated receiver respectively ; and

- Figures 3 and 4 are graphs respectively showing the constellation and spectrum of a digital modulator according to the invention.

The modulator shown in Figure 1 comprises a unit 10 which converts the successive symbols a_i of a digital stream into phase increments $\Delta\phi$. The successive phase increments $\Delta\phi$ produced by the unit 10 are accumulated by a summator 11. The unit 10 may merely consist of a register containing the possible values of the phase increments $\Delta\phi$ and addressed by the current value of the symbol a_i .

In the embodiment of Figure 1, the symbol stream a_i may correspond either to a bit stream b_i or another bit stream c_i having with a lower transmission rate, processed by a redundancy encoder 12. The bit streams b_i and c_i derive from digital sources such as speech encoders, data sources, etc., generally with error correction coding applied. If the encoder 12 is used, the modulator of Figure 1 operates in accordance with a coded modulation (see G. Ungerboeck "Channel coding with multi-level/phase signals", IEEE Trans. on Information Theory, Vol. IT-28, No. 1, January 1982, pages 55-67). The unit 10, summator 11 and encoder 12 are timed by a clock signal CKS at the frequency $1/T_s$ of the symbols a_i .

The summator 11 stores a integer digital value k representing an accumulated phase. This accumulated phase is thus stored as whole multiples of a sub-multiple of π , that is to say in the general form $(k/P)\pi$. For each cycle of the clock CKS, the accumulated phase is incremented by a value $\Delta\phi$ depending on the current symbol a_i . If each symbol represents m bits, $M = 2^m$ different values of the increment may be added in each cycle. These M values are chosen so that the set of the increments is symmetrical with respect to the value 0 so that the spectrum is symmetrical. Values of k of



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the type $k = k' \times K$ will typically be used, where K/P represents the modulation index, and $k' = -M+1, -M+3, \dots, -1, 1, \dots, M-3$ or $M-1$. This choice of equally distributed increments is not the only one possible. For example, $k' = -7, -3, 3$ or 7 could also be used if $m=2$. $\Delta\phi_{\max} = (k_{\max}/P)\pi$ designates the maximum value of the phase increment $\Delta\phi$.

The accumulated phase is fed to a digital filter 15, referred to as phase filter, whose sampling frequency $1/T_e$, set by a clock signal CKE, is higher than the frequency $1/T_s$ of the symbols a_i (generally a multiple of that frequency).

The output signal from the phase filter 15 is a modulating phase Ψ , which a unit 16 converts into a complex signal, i.e. into two real signals, one (I) representing the real component of the complex signal, and the other (Q) representing the imaginary component.

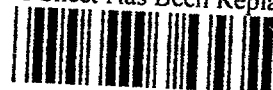
That complex signal has a constant modulus, and an argument equal to the modulating phase Ψ . In other words, $I = \cos\Psi$ and $Q = \sin\Psi$. The unit 16 may merely consist of two read-only memory arrays addressed by the output of the filter 15 at every cycle of clock CKE.

The complex signal is filtered by a digital filter which, in the embodiment shown, consists of two identical filters 17 which respectively filter the components I and Q.

Two digital-analog converters 18 convert the output signals of the two filters 17 into analog form. The two resulting analog signals are fed to low-pass filters 19 in order to eliminate spectral aliasing components. Using respective mixers 21, two quadrature radio waveforms at the carrier frequency, deriving from a local oscillator 20, are mixed with the signals deriving from the anti-aliasing filters 19. The two waveforms thus modulated are combined by a summator 22 whose output is fed to the power amplifier 23 of the transmitter.

If the amplifier 23 were linearised by pre-distortion, it would be necessary to correct the filtered components I and Q, between the filters 17 and converters 18, before converting them into analog form.

The receiver shown in Figure 2 includes a low-noise amplifier 30 which amplifies the signal picked up by the antenna. Its output is converted to



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an intermediate frequency using a mixer 31. A band-pass filter 32 processes the intermediate frequency signal which is then amplified further by an amplifier 33. Two other mixers 34 provide baseband conversion by mixing with two quadrature waveforms. The two quadrature analog components
5 deriving from the mixers 34 are fed to identical low-pass filters 35, then converted into digital form by analog-digital converters 36. The digital components I' and Q' deriving from the converters 36 are fed to a channel demodulator 37.

The demodulator 37 carries out demodulating operations
10 corresponding to the incomplete modulator consisting of the components 10, 11, 15, 16 and (if necessary) 12 of the transmitter shown in Figure 1. Since this incomplete modulator essentially performs continuous phase modulation (CPM), the demodulator 37 may take the form of a conventional CPM demodulator. It may, for example, be based on a demodulation trellis in order
15 to apply the Viterbi algorithm. The demodulator 37 delivers estimates $\hat{b}_i$ or $\hat{c}_i$ of the bits b_i or c_i fed to the modulator.

Advantageously, the demodulator 37 may include two trellis. It uses either one of the trellis, depending on whether the encoder 12 is used at the transmitter or not. The first trellis includes modulation states. In principle,
20 the number of these states is $M^{L-1} \times P$, where L is the memory of the phase filter 15 expressed in number of samples, M is the number of points on the constellation, and P is the denominator of the modulation index. However, it is generally possible to considerably reduce the number of states of the demodulation trellis without adversely affecting the quality of reception
25 significantly. The second one of the trellis further includes the coding states of the redundancy encoder 12, in accordance with the principle of coded modulations. This second trellis is employed if the encoder 12 is used at the transmitter.

In the modulator design, the values of the phase increments $\Delta\phi$
30 are first chosen, as indicated hereabove. The filter 17 which processes the components I and Q and determines the spectral characteristics of the resulting signal is then constructed. The characteristics of this filter 17 must



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be as close as possible to those of the receiving filter consisting of the combination of filters 32 and 35.

A advantageous form of the digital filters 17, used to process the components I and Q, is a filter with a finite impulse response selected to most
5 closely fit a time characteristic of the form:

$$f(t) = \text{Sinc}(\alpha t/T_s) \cdot \text{Sinc}(\beta t/T_s) \cdot e^{-(\pi\gamma t/T_s)^2} \quad (1)$$

where T_s is the duration of a symbol a_i , and $\text{Sinc}()$ is the cardinal sine function. The approximation can be made by choosing the real coefficients α , β and γ . This provides digital filters whose restriction to a finite length is as accurate as
10 possible by virtue of the fast decay of the Gaussian function. The secondary lobes caused by the limitation of the digital filter length are thus minimised.

The following step consists in defining the phase filter 15. The characteristics of this filter 15 are closely related to those of the digital filter 17. A heuristic method is given hereunder based on the following mathematical
15 property: the energy of a complex function $e^{jm(t)}$ with unitary modulus is maximum in a filter whose spectral power template is the Fourier transform of a function $h(t)$ (in other words, that energy is minimum outside the filter) if it satisfies the following equation:

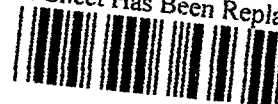
$$\int h(u-t) \cdot e^{jm(t)} dt = \lambda(u) \cdot e^{jm(u)}$$

20 where $\lambda(u)$ is a real function.

The following algorithm is used to define the phase filter 15:

1) A power template filter is selected, i.e. a function $h(t)$ whose Fourier transform represents the required spectral template. A filter identical to the one selected as the I-Q filter is typically chosen. Other choices are
25 obviously possible. In general, it is preferable to use a filter whose digital implementation with a fairly short finite impulse response is possible.

2) A function Φ_0 equal to 0 where $t < 0$, equal to $\Delta\phi_{\max} t/T_s$ where $0 \leq t \leq T_s$ and equal to $\Delta\phi_{\max}$ where $t > T_s$ is used as a first approximation of the phase change function, i.e. of the function available at the output of the phase
30 filter when the maximum phase increment $\Delta\phi_{\max}$ is fed to the accumulator 11. Other approximations using continuous functions equal to 0 where $t < 0$ and



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$\Delta\phi_{\max}$ where $t > T_s$ could be used.

3) A function Φ_n is calculated iteratively using the following formula:

$$\Phi_{n+1}(u) = \frac{\int h(u-t) \cdot \Phi_n(t) dt}{\left| \int h(u-t) \cdot \Phi_n(t) dt \right|}$$

5 4) The n^{th} approximation of the impulse response of the phase filter, which is equal to the derivative of the function Φ_n , is calculated. An approximation of this derived function can also be made using an approximate analytical formula in order to facilitate subsequent calculations. The analytical formula may be as follows:

$$10 \quad g(t) = \text{Sinc}(\alpha' t/T_s) \cdot \text{Sinc}(\beta' t/T_s) \cdot e^{-(\gamma' t/T_s)^2} \quad (2)$$

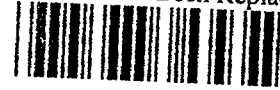
where α' , β' and γ' are real coefficients.

5) For one of these approximations (for example $n = 2$ or 3), the characteristics of the modulation are evaluated with respect to the criteria of interference power in adjacent frequency channel, variation in amplitude and
15 noise immunity. If the approximation is unsuitable, the calculations 1) to 4) are repeated by modifying the values of the phase increments, and/or by modifying the shape of the I-Q filter 17, and/or by modifying the shape of the filter referred to under 1), and/or by modifying the approximation of the phase filter obtained using the algorithm.

20 The retained phase filter is then implemented as a finite impulse response digital filter.

The phase filter 15 of the modulator of Figure 1 could be replaced by a bank of phase filters selected in accordance with the origin of the symbol stream a_i . This could provide a phase filter 15 optimised for cases
25 where the redundancy encoder 12 is not used, and another one optimised for cases where the encoder 12 is used.

In a particular embodiment of the invention, the duration T_s of a symbol a_i is 125 μs . The number of bits per symbol is 2, the phase increments being $-\pi$, $-\pi/3$, $\pi/3$ or π . The bit rate is then 16 kbit/s. The spectral
30 specifications are those of the ETSI standard 300-113. The finite impulse



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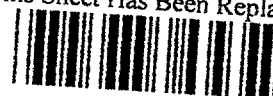
response of the phase filter 15 has a length of 4 symbols and the form (2) with $\alpha' = 0.77$, $\beta' = 0.5$ and $\gamma' = 0$. The finite impulse response of the I-Q filter 17 has a length of 8 symbols and the form (1) with $\alpha = 1.6$, $\beta = 0.1$ and $\gamma = 0.12$. The values specified for the filter parameters could be replaced by values of the same order.

The constellation corresponding to that modulation is shown in Figure 3. A very small variation in amplitude will be noted, since the ratio between the maximum instantaneous power and the average power is only 1.2 dB, whereas the ratio between the maximum and minimum instantaneous powers is less than 2.4 dB. Because of these characteristics, the modulation can be used with weakly linearised power amplifiers, easy to adjust and having an efficiency very close to that of saturated amplifiers.

The spectrum is shown in Figure 4. It can be noted that the level of adjacent channel interference is very low and compatible with the most demanding standards.

The noise immunity performance is excellent since, with a channel affected by Gaussian white noise, an error rate of 1% is noted for a signal-to-noise ratio E_b/N_0 of 5.5 dB in the case of stationary stations, whereas the same error rate of 1% is obtained for a signal-to-noise ratio E_b/N_0 of 16 dB in a dynamic case (speed of 70 km/h and carrier at 400 MHz). These error rates are obtained using simple conventional demodulators (Figure 2), i.e. trellis demodulators with a very small number of states. A trellis with only three states can be used in the example shown.

In the same embodiment of the invention, the optional redundancy encoder 12 permits the implementation of a coded modulation. Here, the redundancy coding is a convolutional coding of rate 1/2, the bit rate then being 8 kbit/s. The filter values are identical, and a trellis with only four states may be used at the demodulator. The coding gain is of the order of 2.5 dB, and an error rate of 1% is noted with $E_b/N_0 = 3.4$ dB on a channel affected by Gaussian white noise in a stationary case.



CLAIMS

1. A digital modulation process wherein the successive symbols (a;) of a digital stream are converted into phase increments ($\Delta\phi$), an accumulated phase is obtained by adding the successive phase increments, a modulating phase (Ψ) is obtained by filtering the accumulated phase, a complex signal is produced whose argument represents the modulating phase, two phase quadrature radio waveforms are respectively modulated on the basis of said complex signal, and a radio signal resulting from a combination of the two modulated waveforms is transmitted, characterised in that said complex signal is digitally filtered and in that the digital signals obtained from the real and imaginary components (I,Q) of the digitally filtered complex signal are converted into analog form before being respectively subjected to anti-aliasing analog filtering and then mixed with the two radio waveforms.

2. A process in accordance with claim 1, wherein the digital filtering of said complex signal consists of two identical filtering operations on the real and imaginary components thereof (I,Q).

3. A process in accordance with claim 2, wherein the digital filtering of the real or imaginary component of the complex signal has a finite impulse response corresponding to a time characteristic having the form:

$$f(t) = \text{Sinc}(\alpha t/T_s) \cdot \text{Sinc}(\beta t/T_s) \cdot e^{-(\gamma t/T_s)^2},$$

where T_s is the duration of a symbol of the bit stream, α , β and γ are real coefficients, and $\text{Sinc}()$ is the cardinal sine function.

4. A process in accordance with claim 3, wherein the filtering of the accumulated phase has a finite impulse response corresponding to a time characteristic having the form:

$$g(t) = \text{Sinc}(\alpha' t/T_s) \cdot \text{Sinc}(\beta' t/T_s) \cdot e^{-(\gamma' t/T_s)^2},$$

where α' , β' and γ' are real coefficients.





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5. A digital modulator comprising: means (10) for converting the successive symbols (a_i) of a digital stream into phase increments ($\Delta\phi$), a summator (11) which accumulates the successive phase increments to produce an accumulated phase, a phase filter (15) receiving the accumulated phase and producing a modulating phase (Ψ), means (16) for producing a complex signal whose argument represents the modulating phase, and a modulator for respectively modulating two phase quadrature radio waveforms on the basis of said complex signal, and for transmitting a radio signal resulting from a combination of the two modulated waveforms, characterised in that the modulator comprises a digital filter (17) to which said complex signal is applied, analog-digital converters (18) respectively processing the digital signals obtained from the real and imaginary components (I,Q) of the digitally filtered complex signal, anti-aliasing analog filters (19) receiving the output signals from the analog-digital converters, and two mixers (21), each receiving one of the two radio waveforms and the output signal from one of the two anti-aliasing filters.

6. A modulator in accordance with claim 5, wherein the digital filter, to which said complex signal is applied consists of two identical filters (17) receiving respectively the real and imaginary components (I,Q) of the complex signal.

7. A modulator in accordance with claim 6, wherein the digital filter (17) processing the complex signal has a finite impulse response corresponding to a time characteristic having the form:

$$f(t) = \text{Sinc}(\alpha t/T_s) \cdot \text{Sinc}(\beta t/T_s) \cdot e^{-(\pi\gamma t/T_s)^2},$$

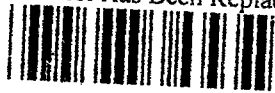
25 where T_s is the duration of a symbol a_i of the bit stream, α , β and γ are real coefficients, and $\text{Sinc}()$ is the cardinal sine function.

8. A modulator in accordance with claim 7, wherein the phase filter (15) has a finite impulse response corresponding to a time characteristic having the form:

$$g(t) = \text{Sinc}(\alpha' t/T_s) \cdot \text{Sinc}(\beta' t/T_s) \cdot e^{-(\pi\gamma' t/T_s)^2},$$

30





where α' , β' and γ' are real coefficients.

9. A modulator in accordance with claim 8, wherein $T_s = 125\mu s$, each symbol (a_i) of the bit stream consists of two bits, the phase increments ($\Delta\phi$) are $-\pi$, $-\pi/3$, $\pi/3$ or π , and $\alpha \approx 1.6$, $\beta \approx 0.1$, $\gamma \approx 0.12$, $\alpha' \approx 0.77$, $\beta' \approx 0.5$ and
- 5 $\gamma' \approx 0$.

