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The invention relates to a method and an apparatus for the digital modulation of a carrier wave in a form which can be represented by two components, an in-phase I component and an in-quadrature component Q. The invention finds a particularly important, although not exclusive application in the modulation with minimum displacement, preceded by Gaussian filtration, currently referred to as GMSK for short and presently proposed for sound transmission to mobile phones and voice transmission between mobile phones (radiotelephony). A description of GMSK modulation can be found in various documents, more particularly by K. Murota "GMSK modulation for digital mobile radiotelephony", IEEE Transactions on Communications, Vol. Com. 29, No. 7, July 1981, pages 1044-1050.

Conventionally GMSK modulation requires that the information to be transmitted be put in analogue form, using digital-to analogue converters. Such converters are also required for other in-phase and quadrature modulation, such as the so-called "MAQ 4 phases" modulation. It is more particularly an aim of the invention to obviate the need for conventional converters, which are expensive components difficult to integrate, while at the same time obtaining a high spectral purity of the signal to be transmitted - i.e., a low noise density at the base of the carrier wave, and an inherent equilibrium of the I and Q paths.

The invention advantageously uses digital signal $\Sigma \Delta$ conversion which has hitherto been used on the baseband. Examples of $\Sigma \Delta$ converters can be found in various documents, more particularly in the article by James Candy, "A use of double integration sigma-delta modulation", IEEE Transaction on Communications, Vol. Com. 33, No. 3,

March 1985, pages 249-258.

To facilitate the explanation of the invention, it may be useful to recall the form of a second order $\Sigma \Delta$ converter, which lends itself to certain variants. Its make up is shown in Fig 1.

Binary output signal Y is obtained from successive samples of input signal X by processing in an adder 10 associated with an external loop, a subtracter 12 associated with an internal loop and a quantization threshold circuit 14 supplying the binary signal Y at its output.

Quantization takes the form of the addition of a quantization error E:

$$Y = S + E,$$

S being the signal at the input of the threshold circuit.

The input signal S of the quantizer is subtracted from the signal quantised in a subtracter 16 whose output passes through an element 18 giving a delay D equal to the sampling period of the input signal. The delayed signal is applied to the subtracting input of the adder 12 after multiplication by 2 - i.e., shifting by one bit. This same signal, delayed by a fresh period D in a delay element 20 is applied to the input of the adder 10.

Such a converter has a transfer function:

$$H(z) = (1-z^{-1})^2 \quad (1)$$

and the quantised output signal Y can be represented, still using the z transforms, by:

$$Y(z) = X(z) + H(z).E(z).$$

When the samples are digital and occur at the frequency $1/T$ in the form of successive words of m bits, the adders 10 and 12 are formed by accumulators having sufficient capacity to obviate the risk of overflow; the

delay elements 18 and 20 are formed by registers having a number of flip flops sufficient for storing the maximum possible value of the quantization error.

5 B. Ekelund et al : "A VLSI sigma delta waveform generator for a quadrature type CPM transmitter", 1987
IEEE Inter. Symposium on Circuits and Systems, Vol 2
IEEE, New York pages 475-478" discloses a wave gener-
ator... for an transmitter with modulation on carrier
10 waves in quadrature. The assembly described is tradi-
tional: the signal is put into analogue form and applied
to analogue carrier waves in quadrature. That apparatus
has the disadvantages and limitations inherent in analo-
gue operation. It is very difficult to obtain two dis-
tinct carrier waves which are really in quadrature and
15 to produce a balanced mix and more generally to balance
the two paths. Moreover, an analogue circuit is poorly
adapted to integration.

A method according to one aspect of the invention
which enables these faults to be obviated is set forth
20 in Claim 1.

One particular aspect of the invention is defined
in Claim 2, using a $\Sigma \Delta$ conversion on the two components
I and Q obtained by GMSK modulation (or more generally
by modulation producing such components) prior to the
25 multiplexing of said components and the formation of the
carrier wave.

It must be remembered that $\Sigma \Delta$ converters are al-
ready used to perform analogue - to - digital or digital
- to - analogue conversion; with a second order conver-
30 ter whose filter has a double zero at the frequency
 $\omega_0 = 0$, the power of the quantization noise E is mini-
mised throughout the band of the required signal: the
resolution of the converter is optimum.

For reasons of hardware or software implementation,
35 the envelope of the components applied to the $\Sigma \Delta$ conver-

ters must be temporarily truncated, so as to reduce the size of the memory required for storing wave shapes. In practice, when GMSK modulation and an oversampling rate of 8 are used, truncation lasting four times the period of bits generally gives satisfactory results.

Often the objective is not a high resolution after GMSK modulation, but a satisfactory spectral purity of the signal emitted after GMSK modulation. Because of the outline of the curve of variation of spectral power density as a function of frequency for a GMSK signal, such purity is appreciably enhanced if a modified second order $\Sigma \Delta$ conversion is used to obtain a transfer function $H(z)$ of the kind:

$$H(z) = 1 - 2 \cos 2\pi\nu_0 \cdot z^{-1} + z^{-2} \quad (2)$$

which has zeros at the frequencies $+\nu_0$ and $-\nu_0$. The formula (2) amounts to case (1) hereinbefore, if $\nu_0 = 0$.

A suitable choice of the frequency of ν_0 reduces the spectral power density of the noise at the base of the carrier wave to such a point that it considerably exceeds a mere compensation of the degradation caused by the displacement of the zero of the signal to noise ratio in the useful band of the signal.

As a rule, the implementation of the above transfer function (2) would be inconvenient, but it can be simplified by synthesising the factor $\cos 2\pi\nu_0 T$ in a polynomial form, such that $1-(1/2)^n$ with a suitable value of n which depends on the form of the typical spectrum of the signal originating from the GMSK modulation.

Another way of enhancing spectral purity, in this case by removing the energy of the noise further in frequency, would be to adopt a $\Sigma \Delta$ converter of order higher than second order.

The invention also proposes an apparatus for the digital modulation of a carrier wave by two components, in-phase and quadrature, from digital samples comprising

a GMSK generator whose outputs, in-phase and in-quadrature are each applied to a path comprising a $\Sigma \Delta$ generator and a modulator which receives the carrier wave at half the frequency of the frequency of the output bits of the $\Sigma \Delta$ converter, the two paths being applied to a mixer.

The invention will be better understood from the following description of the exemplary non-limiting embodiments thereof, with reference to the accompanying drawings, in which:-

Figure 1, which has already been mentioned, shows the basic architecture of a second order $\Sigma \Delta$ converter,

Figure 2 is a skeleton diagram of a GMSK digital modulator with $\Sigma \Delta$ conversion, forming an embodiment of the invention;

Figure 3 is a diagram showing the wave shapes which appear in Fig. 2 at the locations having the same reference to that borne by the lines in Fig 3;

Figure 4 is a diagram showing a possible architecture of the mixer of Fig. 2;

Figure 5 shows the outline of the signals at the input and output of the mixer of Fig.4; and

Figure 6 shows the outline of the spectral power density obtained by the use of a modulator as shown diagrammatically in Figs. 1 and 2.

The modulator, shown diagrammatically in Fig. 2 is adapted to produce a GMSK output signal 24 from a binary modulating signal of bit frequency f_b applied to its input 22.

For this purpose, the modulator shown in Fig. 2 comprises a GMSK signal generator 26 producing output wave shapes X_I and X_Q in the form of digital samples each encoded on a number of bits and a sign bit.

The output signals X_I and X_Q of the generator 26 are applied to two paths each comprising a second order

$\Sigma \Delta$ converter 28 which can have the architecture shown in Fig. 1 and which supplies a single output bit Y_I or Y_Q for each input sample X_I or X_Q . Each path also comprises a modulator 30 receiving a carrier wave having half the frequency of the frequency of the output bits of the $\Sigma \Delta$ converter. The two paths are then mixed in a mixer 32 and the resulting modulated carrier wave is applied to a filter 34 whose role will be explained hereinafter.

The components are timed by signals originating from dividers supplied by a base clock 36 formed by a quartz oscillator.

The various components will be described successively.

The GMSK generator 26 may be constituted by a frequency modulation index $\frac{1}{2}$ which receives the modulated signal constituted by a binary signal at a bit rate of f_b and yields two signals X output encoded on 11 bits and 1 signal bit. The generator 26 comprises a shift register input 38 whose outputs are connected in parallel to a read only memory 40 in the forms I and Q , in the form of cosines and sines of the phase at successive instants $nT + mT_s$, where T is the bit time, $T_s = T/r$ and m is an input number between 0 and $r-1$.

The oversampling rate r forms a compromise between a high value of r , which would call for memory of considerable capacity, and a low value, which would spoil the output spectrum.

By way of example, an apparatus adapted for voice communications was constructed wherein a rate f_b of 8 kbs was adopted, corresponding to a bit period $T = 125 \mu s$. In that case oversampling factor r of 25 gave satisfactory results for obtaining a GMSK signal of frequency modulation index $1/2$, using Gaussian filtration with a cutoff frequency B of 3 dB, such that $BT = 0.25$, with a truncation of the Gaussian line not exceeding $5T$.

For the performance of oversampling in this case, the read only memory 40 has an addressing counter 42 operating at a frequency of $25 f_b$. This frequency is supplied by clock 36 at 40 MHz and two dividers in cascade 44, which reduces the frequency to 20 MHz, and 46, which supplies a frequency of 200 KHz. The counter 42 can be formed by five flip flops in cascade; its carry output supplies the frequency f_b at the shift input of register 38, which receives the bits arriving at input 22. Since truncation is performed on five periods, the register comprises five flip flops in cascade.

The oversampling of the modulated signal in the baseband results in the presence in the resulting spectrum of spectra relocated to frequencies which are multiples of $r \cdot f_b$ and which appear at the input of the $\Sigma \Delta$ converters and are allowed for in the selection of the pulse response of the Gaussian filter. By way of example, satisfactory results have been obtained in the case of the above characteristics with a Gaussian line limited to the interval $(-2.5 T ; +2.5 T)$ on condition that practically the whole energy of the nominal Gaussian line is collected on the truncated Gaussian line.

The GMSK modulator will not be further described, since embodiments thereof are to be found in the aforementioned documents.

Each of the $\Sigma \Delta$ converters 28 can have the basic architecture illustrated in Fig. 1. The quantization scale is selected remembering that the input signal X cannot exceed the values representing the -1 and +1 level at the output without the risk of divergence. Consequently, the values representing the -1 and +1 levels of Y are taken to be equal to -2 048 and +2 048 respectively when X is encoded on 11 bits and 1 sign bit.

The peak values of X must be lower than 2 048 because:

values too close to $\pm 2\ 048$ induce a density of -1 or +1 bits which is too close to unity at the output of the converter 28;

too low peak values increase the quantization noise.

A satisfactory compromise seems to be to adopt peak values equal $\pm 0,8 \times 2\ 048$.

10 A study of the configuration of Fig.1 shows that on condition that the registers 18 and 20 are initialised to 0, it is sufficient to encode the information in the $\Sigma \Delta$ converter on 15 bits and 1 sign bit.

15 The timing frequency of each $\Sigma \Delta$ converter 28 is bound up with the selection of the carrier wave applied to the modulators 30. The carrier wave must be able to receive 1 bit originating from each path at each of its half periods. In that case each $\Sigma \Delta$ converter is timed at 20 MHz by divider 44 if the carrier wave is 10 MHz.

20 The modulation of the carrier wave frequency as shown diagrammatically in Fig. 2 can be considered to be of a pseudo-analogue kind. Each $\Sigma \Delta$ 28 is applied to the input of a multiplier 30 which also receives a M_I or M_Q ternary wave at the fundamental frequency of 10 MHz, the two waves being in quadrature. To this end the modulator shown in Fig. 1 comprises a generator 46 which is applied directly to the multiplier 30 of the path I and is applied to multiplier 30 of path Q via a status phase shifter 48.

30 The wave forms at the input of the multipliers 30 are shown in Fig.3.

The GMSK signal obtained by the summation of the outputs of the multipliers 30 is formed by an NRZ train of 40 Mbs, sampled at 20 MHz. Summing integrator 32 can simply take the form of two OR EXCLUSIVE gates 50 and 52

and a multiplexer 54 (Fig. 4). Figure 5 shows an example of shaping. The output sequences are of the kind Y_{nI} ; Y_{nQ} ; $Y_{(n+1)I}$; $Y_{(n+1)Q}$; $Y_{(n+2)I}$; etc.

5 If the number of encoding bits is that given here-
inbefore, each converter 28 can take the form of a 16
bit adder 10, two 16 bit subtractors 12 and 16 and three
16 bit registers forming the delay elements 18 and 20
and the multiplier by two.

10 Lastly, the modulator shown in Figure 2 comprises
an output bandpass filter 34 whose center frequency is
that of the carrier wave frequency - i.e. 10 MHz, and
whose role is to eliminate the spectral components due
to the quantization noise (twice the width at 3 dB).

15 Another embodiment of the invention can be directly
applied to the characteristic of the PAN EUROPEAN radio-
telephony system. That system provides a bit frequency
 f_b equal to $(13/48)$ MHz, with a channel spacing of 200
KHz. For that system a GMSK modulation installation has
20 been envisaged which comprises digital-to-analogue
converters, whose disadvantages were pointed out herein-
before. The modulating signal is formed by normal bursts
of 148 bits or short bursts, which can be completed by
digits (1) so as to return to the former case.

25 A GMSK generator of the kind shown in Fig. 2 can
also be used, but with an oversampling rate of 8 and and
truncation at $4T$ associated with $\Sigma \Delta$ converters modified
to improve the spectral purity of the signal emitted, so
as to allow the fact that the oversampling rate is much
30 higher than in the preceding case.

To better define the GMSK generator to be used, it
may be useful to recall the complex envelope correspond-
ing to MSK modulation in general. The output signal
 $X(t)$ obtained by this modulation can be represented by:

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$$X(t) = \exp [i(\pi/2) \sum a_n \cdot q(t-nT)] \quad (3)$$

where T is the bit period, and

$$q(t) = \int_{-\infty}^t g(t) \cdot dt$$

with $g(t) = \{1/T \text{ for } 0 \leq t \leq T$
 $\{\text{zero elsewhere}$

5 GMSK modulation differs from MSK-type modulation in that $q(t)$ is generated by a function $g^*(t)$ the form:

$$g^*(t) = h(t) \cdot g(t) \quad (4)$$

where $h(t) = [\sqrt{2} \delta T]^{-1} \cdot \exp(-t^2/2\delta^2 T^2)$

with $\delta = \sqrt{(n(2))}/2\pi BT$

10 $BT = 0.3$ in the case of the Pan-European system.

As was indicated hereinbefore reasons of practical implementation oblige the function $q(t)$ to be truncated. A truncation with a duration $4T$ is sufficient to guarantee satisfactory spectral purity. To this end the Gaussian line can be truncated at $3T$, $q(t)$ can be truncated directly, or $g(t)$ can be truncated at $4T$ prior to integration.

The latter method is advantageous. It modifies the integral of $g^*(t)$, which is no longer strictly equal to 1, something which means that $q(t)$ must be renormalised. This raises no difficulty.

As in the case of the modulator shown in Fig. 2, X can be represented by a word of 11 bits and 1 sign bit, representing values lying between +1 640 and -1 640. However, since the modulating bits take the form of bursts, the output signals must be given an envelope whose level is constant over all the data bits, but which is prolonged by an ascending ramp and a descending ramp, each corresponding, for example, to three bit periods. The oversampling rate being 8, the output samples X_I and X_Q appear with a bit frequency of 26 MHz.

Each of the $\Sigma \Delta$ converters can have the make up shown in Fig. 1, except that the multiplied is provided, to make the transfer function $H(z)$:

$$35 \quad H(z) = 1 - [2 - (1/2)^7]z^{-1} + z^{-2} \quad (2a)$$

This function simulates the formula (2) and corresponds approximately $\nu_0 = 365.9$ MHz. To implement formula (2a) the multiplier is adapted to perform a multiplication by 2 (in the form of a shift), a multiplication by 2^{-7} (shift of 7 bits) and a subtraction.

The effect of the modification of the converter appears in Fig. 6 which shows in continuous lines the spectral power density of the required signal (curve 56) and that of the $\Sigma\Delta$ noise and sampling (curve 58) in the case of GMSK spectrum with $BT = 0.3$ and a second order $\Sigma\Delta$ converter. The resulting curve 60 is modified and takes the form 62 when the $\Sigma\Delta$ converters correspond to formula 2(a) with $\nu_0 = 365.9$ MHz; it can be seen that the energy of the noise is moved towards higher frequencies.

Formula (2a) introduce a quantization noise and a bias of -0.5 in relation to a division by 128, but these have no perceptive effect.

The carrier wave frequency can be modulated as in the case of Fig 2. except that the trains Y_I and Y_Q at 26 Mbs/sec are converted into a single NRZ train at 52 Mbs/sec.

As indicated hereinbefore, the invention is not limited solely to the case of GMSK modulation, but can more particularly be applied to all continuous phase modulations with full or partial response. Whatever the embodiment adopted may be, the symmetry of the carrier wave ensures a perfect balance, unachievable with converters which cannot be absolutely identical.

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CLAIMS

1. A method of modulating a carrier wave by two components, in-phase and in-quadrature, starting from digital signals, characterized in that the modulation components, in-phase and in-quadrature, are formed by two sequences of binary elements of pre-determined frequency, and said sequences of binary elements are applied to a symmetrical carrier wave having three levels +1, 0 and -1, at half the frequency of the frequency of the binary elements.
2. A method according to Claim 1, characterized in that a time multiplexing is obtained which produces a digitally modulated carrier wave with a modulation frequency twice the frequency of each sequence, by the application of the two sequences of two components of said carrier wave in-quadrature and the addition of the ternary components in-quadrature modulated by the sequences, alternate elements of each sequence being inverted.
3. A method according to Claim 1 or 2, characterized in that said sequences of binary elements are obtained by the $\Sigma \Delta$ conversion of two trains of digital signals, in-quadrature, obtained from samples of a modulating input signal.
4. A method according to Claim 3, characterized in that the $\Sigma \Delta$ conversion is performed with a transfer function $H(z)$ simulating:

$$H(z) = 1 - 2 \cos(2\pi\nu_0) \cdot z^{-1} + z^{-2}$$

where ν_0 is a frequency selected to reduce the spectral power density of the noise at the base of the carrier wave.

5. A method according to Claim 3, characterized in that the $\Sigma \Delta$ conversion is performed with a transfer function $H(z)$

$$H(z) = 1 - [2 - (2)^{-n}]z^{-1} + z^2$$



where $7 < n < 9$.

6. A method according to any of Claims 1 to 5, characterized in that the components are created by GMSK modulation from a modulating input signal, followed by
5 conversion into sequences of binary elements.

7. A method according to Claim 6, characterized in that the frequency of bits at the GMSK modulation output being 26 MHz, ν_0 is selected equal to 365.9 MHz.

8. A method according to Claim 5 characterized in that
10 the components obtained by GMSK modulation are oversampled at eight times the bit frequency of the input signal.

9. Use of the method according to any of the preceding Claims in a mobile radiotelephony system.

15 10. An apparatus for the digital modulation of a carrier wave by two components, in-phase and in-quadrature, from digital samples, characterized in that it comprises a GMSK generator whose outputs, in-phase and in-quadrature, are each applied to a path comprising a $\Sigma \Delta$
20 generator (28) and a modulator (30) which receives the carrier wave having half the frequency of the frequency of the output bits of the $\Sigma \Delta$ converter, the two paths whose carrier waves are in quadrature, being applied to a mixer (32).

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