TRANSMITTER FOR QAM ENCODED DATA

BACKGROUND

The present invention relates to the transmission of digital information over channels of limited bandwidth, such as radio channels of telephone lines. The digital information can, for example, comprise digitally coded voice.

It is known that Quadrature Amplitude Modulation (QAM) comprises encoding data bits into complex vector signals in which the real and imaginary parts can each take on one of a multiplicity of levels. For example, in 16QAM the real part and the imaginary part can each take on one of the four equispaced values 3, 1, -1 or -3. The 4x4 possible points that arise are called the constellation, and are shown in FIG. 1(a). Modulation in accordance with this technique comprises mapping four-bit data values, designated $B_0B_1B_2B_3$, onto the 16 distinct complex signal values, as shown in the figure.

A prior art 16QAM transmitter for generating the above-described constellation from the four-bit data values is illustrated in FIG. 1(b). A first pair of bits, B_0B_1, is supplied to a first 2-bit Digital to Analog (D/A) convertor 101 in order to determine which of four values the real part will attain. A second pair of bits, B_2B_3, is supplied to a second 2-bit D/A convertor 103 in order to determine which of four values the imaginary part will attain. Each of the D/A convertors 101, 103 supplies its output to a corresponding one of two low-pass filters 105, 107. The function of the low-pass filters 105, 107 is to contain the transmitted spectrum by smoothing transitions from one value to the next when any of the bits $B_0B_1B_2B_3$ are changed. The low-pass filters 105, 107 are preferably Nyquist filters, which have the property that, at regular sampling times after a bit change, the filter output will

exactly attain the value determined by the input bits $B_0B_1B_2B_3$. The smoothed real values 109 are applied to a cosine modulator 113, while the smoothed imaginary values 111 are applied to a sine wave modulator 115. The modulated sine and cosine waves 117, 119 are summed at summation point 121 to form a complex modulated carrier signal 123 which varies in both phase and amplitude. To preserve amplitude variations, the prior art requires that this signal be amplified by a linear power amplifier 125. Due to the Nyquist property of the low-pass filters 105, 107, if the output signal vector 127 is sampled at the proper regular instants, one of 16 complex values will be observed as shown in the grid diagram of FIG. 1(a).

The prior art 16QAM transmitter has the drawback that the linear power amplifier 125 is not efficient, and if it exhibits distortion or non-linearities, then the desired 16 constellation points are not observed in the output signal vector 127. Equally, if the communications channel including the low pass filters 105, 107 is not strictly Nyquist, intersymbol interference (ISI) will prevent the desired constellation points from being observed.

US-A-4 485 357 describes a method of producing a signal that is continuously modulated in both amplitude and phase, such as an analog SSB signal. The method of this patent was described in a paper by Chireix, "High Power Outphasing Modulation, PROC. IRE, vol. 23, No. 11 (1935). This prior art method involves coupling two equal power amplifiers together and driving them with signals that differ in phase angle, by the arccosine of the ratio of the desired amplitude to the peak amplitude. It does not disclose techniques for digitally modulating bits.

US-A-4 737 968 discloses using a phase-locked loop to filter the phase of a principally phase modulated signal.

This filter removes any amplitude modulation that may originally have been present. The phase-filtered signal is then upconverted (in their mixer 84) to the desired transmit frequency.

EP-A-0 382 697 discloses a control voltage generator in a transmitter arrangement for digitally modulated signals. This document shows the use of constant envelope power amplifiers in a QAM transmitter, and the use of OQPSk for encoding data bits.

SUMMARY

According to one aspect of the invention, an apparatus for generating modulated signals from data bits is provided. In one embodiment, Quadrature Amplitude Modulated signals are generated from data bits by using a first Quadrature Phase Shift Keying (QPSK) modulator for encoding a first pair of the data bits into one of four carrier signal phases, thereby producing a first QPSK signal. A second QPSK modulator encodes a second pair of the data bits into one of four carrier signal phases, thereby producing a second QPSK signal. The first QPSK signal is amplified to a first power level, and the second QPSK signal is amplified to a second power level. The first and second amplified signals are then combined to produce a signal in which four data bits are encoded.

According to a second aspect of the invention, a new, spectrally-efficient modulation is disclosed that is termed Offset Quadrature Phase Shift Keying (OQPSK) modulation, in which data bits are encoded alternately into levels of a cosine wave carrier and levels of a sine wave carrier. In one embodiment, data bits are encoded by encoding a first sub-group of the data bits into a real part of a complex signal at an odd instant of a clock, and

by encoding a second sub-group of the data bits into an imaginary part of the complex signal at an even instant of the clock. OQPSK modulation provides the benefit of having all signal transitions being constrained to smooth trajectories around constant radius circles, thereby allowing a better spectral containment when constant envelope power amplifiers are used.

According to a third aspect of the invention, an inventive transmitter is disclosed for Offset Quadrature Amplitude Modulation comprising two or more power amplifiers that amplify Offset QPSK, MSK or GMSK signals formed from first and second pairs of bits of an OQAM symbol. The first QPSK (or MSK or GMSK) signal is amplified to a first power level, and the second QPSK (or MSK or GMSK) signal is amplified to a second power level. The first and second amplified signals are then combined to produce a signal in which four data bits are encoded.

BRIEF DESCRIPTION OF THE DRAWINGS

The objects and advantages of the invention will be understood by reading the following detailed description

in conjunction with the drawings in which:

FIG. 1(a) is a grid diagram of a constellation of complex signal values generated in accordance with a prior art 16QAM transmitter;

FIG. 1(b) is a diagram of a prior art 16QAM transmitter;

FIG. 2 is a block diagram of a transmitter in accordance with one aspect of the invention;

FIGS. 3(a) - 3(e) illustrate signals and constellation points associated with various nodes of one embodiment of the inventive transmitter;

FIG. 4 is a block diagram of a transmitter employing OQPSK modulation to provide Offset 16QAM in accordance with another aspect of the invention; and

FIGS. 5(a) - 5(h) illustrate signals and constellation points associated with various nodes of one embodiment of the inventive Offset 16QAM transmitter.

DETAILED DESCRIPTION

The various features of the invention will now be described with respect to the figures, in which like parts are identified with the same reference characters.

FIG. 2 is a block diagram of a transmitter 200 in accordance with one aspect of the invention. One advantage of the transmitter 200 is its capability of reproducing the 16QAM constellation points even if non-linear amplifiers are employed.

In a preferred embodiment, a first Quadrature Phase Shift Keying (QPSK) modulator 201 receives two information bits, B_0B_1, and modulates these onto a carrier wave in accordance with well-known techniques. That is, the QPSK constellation encodes two bits into one of the four vector values ± 1 $\pm j$ by changing a real part (I or cosine component) between the values +1 and -1 according to a first information bit and an imaginary part (Q or sine

component) between the values +j and -j according to a second information bit. Since all four vectors that can be produced are of the same amplitude (i.e., $\sqrt{2}\approx 1.414$), the output signal from the first QPSK modulator 201 can be faithfully amplified by a constant envelope power amplifier 203. The constant envelope power amplifier 203, as well as others employed when practicing the invention, may alternatively be a power amplifier operated at output saturation, a class-C amplifier, or a class-B amplifier. The I and Q components that appear at the output of the constant envelope power amplifier 203 are illustrated in FIG. 3(a), and the corresponding constellation of vectors is shown in FIG. 3(b).

The two remaining information bits, B_2B_3, are encoded into another QPSK constellation by means of a second QPSK modulator 205. The output signal from the second QPSK modulator 205 is then supplied to a second constant-envelope amplifier 207 which faithfully amplifies that signal to a power level that is half that of the first constant-envelope amplifier 203, so that the amplitudes of the I and Q components will be $\frac{1}{\sqrt{2}}$ times that which are produced by the first constant-envelope amplifier 203. The amplified I and Q components that appear at the output of the second constant-envelope amplifier 207 are illustrated in FIG. 3(c). The constellation points generated at the output of the second constant-envelope amplifier 207 are thus described by $\frac{\pm 1 \pm j}{\sqrt{2}}$, and shown in FIG. 3(d).

The outputs of the first and second constant-envelope

amplifiers 203, 207 are then supplied to respective inputs of means for summing these signals, such as the directional coupler 209 shown in FIG. 2. The directional coupler 209 further scales the lower power signal from the second constant-envelope amplifier 207 by an amount that is $\frac{1}{\sqrt{2}}$ relative to the voltage scaling of the higher power signal from the first constant-envelope amplifier 203. The further scaled signal is then added to the higher power signal. Such scalings that can be achieved using passive, lossless combining networks such as the directional coupler 209 are limited to the values described by k and $\sqrt{1-k^2}$.

To achieve a relative scaling of $\frac{1}{\sqrt{2}}$, the coupling factor k for the lower power signal should be $\frac{1}{\sqrt{3}}$ and then the coupling factor for the higher power signal is:

$$\sqrt{1-k^2} = \sqrt{1-\frac{1}{3}} = \sqrt{\frac{2}{3}} = \frac{\sqrt{2}}{\sqrt{3}}$$

The further relative scaling of $\frac{1}{\sqrt{2}}$ of the lower power signal relative to the higher power signal, combined with its already $\frac{1}{\sqrt{2}}$ signal level relative to the higher

power signal, combines to produce a signal of relative level $\frac{1}{2}$ that of the higher power signal. Thus the higher power signal of ±1 ±j combines with a lower power signal further scaled to $\frac{\pm 1 \pm j}{2}$ to produce the sixteen constellation points with real and imaginary parts each taking on one of the four values ±1.5, ±0.5 that are then further reduced by the overall scaling of $\sqrt{\frac{2}{3}}$ produced by the directional coupler 209. For the constellation points of largest magnitude (∓1.5 ∓1.5j) the output values from the directional coupler are thus $(\mp\sqrt{\frac{3}{2}} \mp\sqrt{\frac{3}{2}}j)$, corresponding to a peak power level of $\left(\sqrt{\frac{3}{2}}\right)^2 + \left(\sqrt{\frac{3}{2}}\right)^2 = 3$. This is equal to the sum of the amplifier powers. Therefore, the coupling arrangement is 100% efficient at the peak power output level. Other distributions of the overall scaling of $\frac{1}{2}$ between the signal modulated with the most significant bits and that of the least significant bits could be used, other than

allocating $\frac{1}{\sqrt{2}}$ to the power amplifier differences and $\frac{1}{\sqrt{2}}$ to the relative coupling. However, the preferred arrangement gives the maximum efficiency at peak power output points. The 16 QAM constellation points, which are illustrated in FIG. 3(e), are thus faithfully reproduced despite the use of constant envelope (non-linear) amplifiers 203, 207.

In the inventive transmitter shown in FIG. 2, smoothing of transitions between constellation points is accomplished within the first and second QPSK modulators 201, 205 either by smoothing I or Q transitions (so called linear filtering) or by smoothing phase transitions from one point to another. Linear filtering causes the signal to deviate from constant amplitude between constellation points, and the non-linear, constant envelope amplifiers 203, 207 will distort these amplitude variations. Nevertheless, if the vectors arrive at their exact values at a proper sampling time, the constellation points will be reached exactly. The distortion between times gives rise to spectral energy broadening into neighboring channels. Nevertheless, the spectral containment of the transmission using the inventive 16 QAM transmitter can be better than the spectral containment of other, prior art modulations suitable for use with constant envelope power amplifiers. Distortion in power amplifiers may also be reduced by use of well-known prior art predistortion techniques such as described in U.S. Patent No. 5,191,597 to Ekelund et al., which is hereby incorporated by reference. In Ekelund et al., a method is shown for compensating for non-linearities in an end amplifier

having a given transfer function H_R, H_ϕ (for amplitude and phase, respectively) and included in a radio transmitter of quadrature type for linear, digital modulation, and in which table look-up units (ST, CT) store the digital sine and cosine values ($I(t,\alpha)$, $Q(t,\alpha)$) of the quadrature components determined by a given signal vector α. According to the method of values of the transfer functions H_R, H_ϕ for the quadrature modulated radio signals $r(t,\alpha)$ are calculated by addressing memory units which store a number of values of H_R and H_ϕ. The sine and cosine values of the addressed values of H_R and H_ϕ are also formed. The thus calculated values are multiplied by the stored digital values in the table look-up units (ST, CT) and by the inverted value of H_R. As a result, new modified values $i(t,\alpha)$, $q(t,\alpha)$) are obtained for the quadrature components, which compensates for the non-linearities in the final amplifier.

It is also possible to smooth transitions between constellation points in a constant amplitude trajectory. For example, the transition between the values 1+j and 1-j in the QPSK constellation produced by the constant envelope amplifier 203 can be achieved by a clockwise movement through 90 degrees around a circle having a radius equal to $\sqrt{2}$. Using QPSK however, transition to a diametrically opposite point can be required, and neither a clockwise nor a counterclockwise rotation through 180 degrees around a constant radius circle provides good spectral containment compared to a transition through the origin, which is a non-constant amplitude transition that may not be handled satisfactorily by the constant envelope amplifiers 203, 207.

According to a further aspect of the invention, a new modulation entitled Offset QAM is provided in which the

diametric transitions of the constituent QPSK signals are prevented by changing real and imaginary parts at alternate time intervals instead of at the same time. This modulation in its most general form does not have the property that a data symbol corresponds to one of a number of constellation points on a grid, but rather has the property that half of the underlying data bits are encoded in the real values of a complex signal attained at, say, odd intervals of a data clock and the other half are encoded in the imaginary values of the complex signal attained at, in this example, even intervals of the clock. Of course, the designation of alternate clock intervals as the "odd" or the "even" intervals is arbitrary, and not meant to be limiting. Continuing, however, with our first example, data is decoded or demodulated by sampling the real part of the signal at odd intervals, at which times the imaginary parts are between imaginary values or otherwise indeterminate, and by sampling the imaginary part at even intervals of time when the real part is indeterminate. No grid of constellation points thus exists, but rather a set of imaginary or horizontal "stripes" at even times and a set of real or vertical stripes at odd times.

Offset QPSK has transitions only between, for example, 1+j and 1-j (i.e., through 90 degrees), and never diametric transitions through 180 degrees, such as from 1+j to -1-j. Thus all transitions can be constrained to trajectories around constant radius circles. This constraint yields constant envelope signals that can be amplified by class-C power amplifiers having high efficiency. The spectral containment of the resultant Offset 16QAM signal is thus improved compared to prior art 16QAM modulations using non-linear amplifiers.

An embodiment of a transmitter 400 employing Offset 16QAM will now be described with respect to FIG. 4. In

this embodiment, a first Offset QPSK modulator 401 receives two information bits, B_0B_1, and modulates these onto a carrier wave in accordance with the inventive techniques described above. That is, one of the bits, say B_0, is encoded by changing a real part (I or cosine component) between the values +1 and -1 at odd intervals of a data clock according to the value of the information bit B_0, and the other bit (B_1 in this example) is encoded by changing an imaginary part (Q or sine component) between the values +j and -j at even intervals of the data clock according to the value of the information bit B_1. These real and imaginary parts are combined within the Offset QPSK modulator 401 and the resulting signal is supplied to a constant envelope amplifier 403. Since all transitions of this signal are constrained to trajectories around constant radius circles, thereby yielding constant envelope signals, these can be faithfully amplified by the constant envelope amplifier 403, which may be a class-C power amplifier having high efficiency. The I and Q components that appear at the output of the constant envelope power amplifier 403 are illustrated in FIG. 5(a). As explained above, this type of modulation does not yield a constellation of point vectors, as in conventional QPSK modulation. Instead, the real components attain values of ±1 at odd clock times when the imaginary components are indeterminate as illustrated in FIG. 5(b), and the imaginary components attain values of ±j at even clock times when the real components are indeterminate, as illustrated in FIG. 5(c).

The two remaining information bits, B_2B_3, are encoded into another Offset QPSK set of vertical and horizontal stripes by means of a second Offset QPSK modulator 405. The output signal from the second Offset QPSK modulator 405 is then supplied to a second constant-envelope amplifier 407 which faithfully amplifies that signal to a

power level that is half that of the first constant-envelope amplifier 403, so that the amplitudes of the I and Q components will be $\frac{1}{\sqrt{2}}$ times that which are produced by the first constant-envelope amplifier 403. The amplified I and Q components that appear at the output of the second constant-envelope amplifier 407 are illustrated in FIG. 5(d). The amplified vertical stripes that appear during odd clock intervals are illustrated in FIG. 5(e), and the amplified horizontal stripes that appear at the output of the second constant-envelope amplifier 407 during even clock intervals are illustrated in FIG. 5(f).

The outputs of the first and second constant-envelope amplifiers 403, 407 are then supplied to respective inputs of means for summing these signals, such as the directional coupler 409 shown in FIG. 4. The directional coupler 409 further scales the lower power signal from the second constant-envelope amplifier 407 by an amount that is $\frac{1}{\sqrt{2}}$ relative to the voltage scaling of the higher power signal from the first constant-envelope amplifier 403. Techniques for achieving this relative scaling in a directional coupler were described above with respect to the embodiment of FIG. 2. The further scaled signal is then added to the higher power signal.

The further relative scaling of $\frac{1}{\sqrt{2}}$ of the lower power signal relative to the higher power signal, combined

with its already $\frac{1}{\sqrt{2}}$ signal level relative to the higher power signal, combines to produce a signal of relative level $\frac{1}{2}$ that of the higher power signal. Thus the higher power signal combines with a lower power signal further scaled to $\frac{\pm 1 \pm j}{2}$ to produce the vertical (real) stripes of FIG. 5(g) during odd clock times, and the horizontal (imaginary) stripes of FIG. 5(h) during even clock times. The vertical stripes may assume any one of the four values ±1.5, ±0.5 which are further reduced by the overall scaling of $\sqrt{\frac{2}{3}}$ produced by the directional coupler 409. Similarly, the horizontal stripes may assume any one of the four values ±1.5j, ±0.5j further reduced by the overall scaling of $\sqrt{\frac{2}{3}}$ produced by the directional coupler 409. By sampling the signal at an odd clock time and also at its corresponding even clock time, one of sixteen different values may be determined.

In another aspect of the invention, the principles outlined above may be extended to higher order QAM constellations having, for example, 64 or 256 points by the addition of a third or fourth constant envelope power amplifier with suitable power scaling and scaled addition using output couplers such that each contributes part of

the output constellation in the binary voltage ratios 1:1/2:1/4... , and so on. Such variations are considered to lie within the scope of this invention as described by the attached claims.

It is well known in the art that Offset QPSK, having transitions of no more than 90 degrees in phase angle, lends itself more readily to be spectrally contained when the vector trajectory is constrained to follow a constant-envelope circle by smoothing the changes in angle. The rate-of-change of phase angle is by definition the instantaneous deviation of the frequency of the signal from its nominal center frequency. If the phase angle changes from one data bit period to the next take place at a constant rate, i.e. by rotating the phase angle through ±90 degrees over one bit interval, the equivalent frequency change is one quarter of a cycle per bit period, or B/4 Hz, where B is the bitrate. Constant-rate phase changes of this kind give rise to a form of constant envelope OQPSK called Minimum Shift Keying (MSK).

In MSK, angle changes occur at a constant smooth rate, but their derivative, instantaneous frequency, changes in an abrupt fashion as the direction of the phase change alters from a clockwise to a counterclockwise rotation. Thus the angle is a continuous function, its derivative (frequency) is also a continuous function (but having abrupt steps), however the derivative of frequency has infinite discontinuities (dirac functions) at the abrupt step points.

The rate at which the spectrum falls off out of the wanted signal bandwidth is 6N dBs per octave, where N is the order of the lowest order derivative to contain discontinuities; thus because the second derivative of the

phase angle contains discontinuities in the case of MSK, the spectrum falls off out-of-band at 12dB per octave.

By further filtering the frequency waveform so that the abrupt steps are replaced by smooth changes in frequency, the discontinuities may be displaced from the second derivative of phase to ever higher orders, thus causing the spectrum to fall off faster. There is a limit to the spectral containment however whenever constant envelope modulation is used, as the actual components of the transmitted signal are proportional to the Sine and Cosine of the phase, which are non-linear functions. It is found empirically that under these conditions, a Gaussian-shaped low-pass filtering of the frequency waveform produces the best spectral containment achievable; this variant of MSK is known as Gaussian-filtered Minimum Shift Keying (GMSK) and is used in the European Digital Cellular system known as the Global System for Mobile communication (GSM). GMSK is in fact a family of modulations described by the parameter BT, the product of the Gaussian filter's -3dB bandwidth and the information bit duration (bit period), T. Smaller values of BT provide tighter spectral containment at the expense of the signal not quite reaching the nominal constellation points before it starts on a new trajectory for the next bit period, a phenomenon known as "partial response", which makes the signal more difficult to decode efficiently. The compromise between spectral containment and the partial response phenomenon is left to the designers of any particular system.

Other frequency-waveform filters such as Nyquist filters may be used, that guarantee that the signal (at least as transmitted if not as received) passes exactly through the nominal constellation points, i.e. it does not exhibit the partial response effect.

Whenever a transition between constellation points has been smoothed by filtering, the shape of the transition (the trajectory) depends not just on the start and end points (the current data bit) but also on previous

and future data bits. The number of consecutive data bits on which the shape of the trajectory depends is equal to the impulse response length of the filter. If this is a finite number of data bit periods L, as when a Finite Impulse Response (FIR) filter is used, then a finite number two-to-the-power L of different trajectory shapes is produced, corresponding to all possible patterns of L binary bits. These waveforms may be precomputed and stored in a look-up table as a series of waveform samples, and recalled based on an address formed using L consecutive data bits. Recalled samples may be digital-to-analog (D/A) converted to generate analog, I,Q modulating waveforms. The D/A convertors may also be eliminated by storing precomputed bit sequences of Delta-Sigma modulation representations of the waveforms, as described in U.S. Patent No. 5,530,722 which is hereby incorporated herein by reference.

It has been explained above how a certain group of modulations comprising at least OQPSK, MSK and GMSK are closely related to each other and represent data by essentially the same constellation points, differing only in the shape of the transitions between constellation points. Any of these modulations may be used in the current invention to encode pairs of data bits, signals encoded with different pairs of data bits then being scaled and added to form the inventive Offset-QAM signal.

The invention has been described with reference to a particular embodiment. However, it will be readily apparent to those skilled in the art that it is possible to embody the invention in specific forms other than those of the preferred embodiment described above. The preferred embodiment is merely illustrative and should not be considered restrictive in any way.

Claims:

1. An apparatus for encoding groups of data bits into a complex signal for transmission, the apparatus comprising:

means (401, 40, 405, 407, 409) for encoding a first sub-group of the data bits (B0, B2) into a real part of a complex signal at an odd instant of a clock; and

means (401, 40, 405, 407, 409) for encoding a second sub-group of the data bits (B1, B3) into an imaginary part of the complex signal at an even instant of the clock.

2. The apparatus of claim 1, wherein:

the means for encoding the first sub-group' comprises means for choosing one of a number of predetermined signal values according to bit polarities of the first sub-group of the data bits; and

the means for encoding the second sub-group comprises means for choosing one of the number of predetermined signal values according to bit polarities of the second sub-group of the data bits.

3. The apparatus of claim 2 , wherein the predetermined signal values are equispaced with respect to one another.

4. A method of encoding groups of data bits into a complex signal for transmission comprising the steps of:

encoding (401, 403, 405, 407, 409) a first sub-group of the data bits (B0, B2) into a real part of a complex signal at an odd instant of a clock; and

encoding (401, 403, 405, 407, 409) a second sub-group of the data bits (B1, B3) into an imaginary part of the complex signal at an even instant of the clock.

5. The method of claim 4, wherein:

the step of encoding the first sub-group comprises choosing one of a number of predetermined signal values according to bit polarities of the first sub-group of the data bits; and

the step of encoding the second sub-group comprises choosing one of the number of predetermined signal values according to bit polarities of the second sub-group of the data bits.

6. The method of claim 5, wherein the predetermined signal values are equispaced with respect to one another.