RECOMMENDATION ITU-R F.763-5*

Data transmission over HF circuits using phase shift keying or quadrature amplitude modulation

(Question ITU-R 145/9)

(1992-1994-1995-1997-1999-2005)

Scope

This Recommendation provides data transmission systems using phase-shift keying (PSK) and quadrature amplitude modulation (QAM) over HF channels. Information is contained in Annex 6 for data rates from 3 200 to 12 800 bit/s.

The ITU Radiocommunication Assembly,

considering

a) that there is an increasing demand for high-rate data transmission;

b) that to meet this need, two types of phase-shift keying (PSK) modems may be used, namely parallel transmission modems using multi-channel voice frequency telegraphy and serial transmission modems using a single sub-carrier;

c) that to compensate for the unfavourable nature of the transmission medium, the following techniques are available for the two types of modems:

– various forms of dual diversity operation including separate single sideband (SSB) emissions or a single independent sideband (ISB) emission;
– error detection and error correction coding combined with time interleaving;
– variable data rate to adapt the system to the channel capacity;

and, for parallel transmission modems only:

– several levels of in-band frequency diversity;
– introduction of guard times between frames to combat multipath propagation and group-delay distortion,

recommends

1 that for data transmission at binary data rates up to 2400 bit/s using frequency-division multiplex (FDM) and PSK systems, the system described in Annex 1 is preferred;

2 that for data transmission at binary data rates up to 3600 bit/s using serial transmission modems, the system described in Annex 2 is preferred;

3 that reference should be made to Annex 3 for additional information concerning generalized PSK;

* This Recommendation should be brought to the attention of Radiocommunication Study Group 8.

4 Annex 4 describes mode/polarization diversity systems to improve the performance of HF PSK systems;

5 that for data transmission at binary rates up to 4 800 bit/s using serial transmission modems, the system is described in Annex 5;

6 that for data transmissions at binary rates from 3 200 to 12 800 bit/s using serial transmission modems, the preferred system characteristics are described in Annex 6.

Annex 1

Data transmission at 2 400/1 200/600/300/150/75 bit/s over HF circuits using multi-channel voice-frequency telegraphy and PSK

1 System description

1.1 A receiving/transmitting terminal of the system consists of:

– a sender and receiver of digital information (e.g. computer);

– a modem, the primary function of which is the conversion of information from digital to analogue form compatible with the input to a radio transmitter and conversion of the analogue information at a radio receiver output into digital data compatible with the digital receiver input.

  This modem also performs various coding functions and effects diversity combination;

– RF receiving and transmitting equipment connected to antennas.

1.2 At the transmit side, the 2 400 bit/s incoming data stream is fed to a serial-to-parallel converter. At 32-bit intervals (i.e. 13.33 ms intervals) the content of this converter is transferred in parallel to a 32-bit memory device, the output of which is connected to a QPSK modulator.

The modem generates in transmission a composite audio signal consisting of a set of 18 tones in the band 300 to 3 000 Hz.

Of these tones, 16 have a spacing of 110 Hz (935 to 2 585 Hz) and are modulated in differentially encoded quaternary phase shift keying (DE-QPSK) mode, each at 75 Bd, thus permitting a data rate of $16 \times 75 \times 2 = 2\,400$ bit/s.

The tone at 605 Hz is used for the correction of end-to-end frequency errors, including any Doppler effect. The tone at 2 915 Hz (or 825 Hz) is used for system synchronization.

The dual diversity combiner can accept inputs either from two receivers operating in space, frequency or polarization diversity mode or from one receiver operating in ISB mode.

When the data rate is a sub-multiple of the transmission speed, various in-band diversity arrangements can be implemented. As an example, a data rate of 1 200 bit/s provides a dual diversity (1 200 × 2), a data rate of 600 bit/s, a quadruple diversity (600 × 4) and so forth, all with a transmission speed of 2 400 bit/s. Utilization of the maximum possible diversity, both in-band and between independent channels, can thus be made according to the data rate selected. Provision is made for 75/150/300/600/1 200 bit/s.

In addition to a choice of coded/uncoded operation, with selectable data rate and diversity mode, this modem also allows setting of the interleaving interval thus providing a flexible communication system as summarized in Table 1.

The transmission signal consists of frames whose duration is 13.33 ms. This includes a time guard (4.2 ms) which is introduced to offset the effects of multipath propagation.

The modem uses two techniques to reduce signal impairments, particularly those caused by impulsive noise and flat fading:

– error correction code;

– time interleaving.

A form of BCH cyclic block code (16,8) is used. The BCH codewords are stored in a memory to be extracted during the interleaving process. Interleaving is obtained by considering:

– the first bit of the last stored word;

– the second bit of the "(m) word stored before";

– the third bit of the "(2 m) word stored before" ...;

– the 16th bit of the "(15 m) word stored before".

TABLE 1

Data rates/modes (independently selectable for transmission and reception)

Data rate (bit/s)	Uncoded modes			Coded modes			
	Diversity modes			Time interleaving Available time spread (transmitter and receiver) (s)	Additional diversity modes		
	In-band	Channel	Total		In-band	Channel	Total
2 400	–	×2	×2				
1 200	×2	×2	×4	0-12.8	–	×2	×2
600	×4	×2	×8	0-25.6	×2	×2	×4
300	×8	×2	×16	0-51.2	×4	×2	×8
150	×16	×2	×32	0-102.5	×8	×2	×16
75				0-205	×16	×2	×32

The interleaving level (m codewords) can be chosen according to the propagation conditions of the radio path from 0 (no interleaving), 1, 2, 4, 8, 16, 32, or 64, corresponding to a data reception delay ranging from a few milliseconds to tens of seconds. As the wrong bits do not belong to the same coded word, a better protection against burst errors is achieved.

In Fig. 1, the performance of the modem with Gaussian distributed noise is given in terms of bit error probability, P_e, as a function of signal-to-noise ratio, S/N, for both with coding and without coding modes, in a 250-3 000 Hz bandwidth.

The effects of coding become prominent at the higher values of S/N.

The curves were obtained with an experimental test set-up in which the modem was fed with a test pattern to produce the audio frequency tones. The output of the modem was summed with Gaussian noise, filtered and applied to the receiving input of another modem from which the test pattern was retrieved at the output. The test pattern was then fed to a data error analyzer to enable the bit-error ratio (BER) to be determined.

Figure 2 indicates the results of a computer simulation of the modem performance in a fading channel.

A fading channel was simulated in which two equi-amplitude paths carry signals separated by a multipath delay of 1 ms and differing in frequency by 1 Hz, in order to obtain fades which ran through the passband rather than remaining at certain fixed frequencies.

From Fig. 2, it can be seen that the performance is improved by using a combination of the various types of diversity (in-band and out-of-band), error correcting codes and interleaving techniques for 600, 1 200 and 2 400 bit/s rates.

The modem is currently in experimental use as part of an HF link between two radio stations located in central and southern Italy, and separated by approximately 800 km (500 miles).

1.3 The RF equipment performs, in transmission, operations relative to channel modulation and produces an emission having suitable radio frequency and power characteristics. Reverse operations, relative to frequency conversion, are carried out in reception so as to obtain the composite audio signal to be conveyed to the modem.

The RF equipment has the following specific characteristics:

– phase jitter: less than 5° for 10 ms time interval (100 samples);

– group delay distortion: 500 µs in transmission, 500 µs in reception;

– intermodulation: 36 dB below peak envelope power.

FIGURE 1

Bit error probability versus *S/N* for various data rates using with coding or without coding modes with in-band diversity for a non-fading channel with Gaussian noise

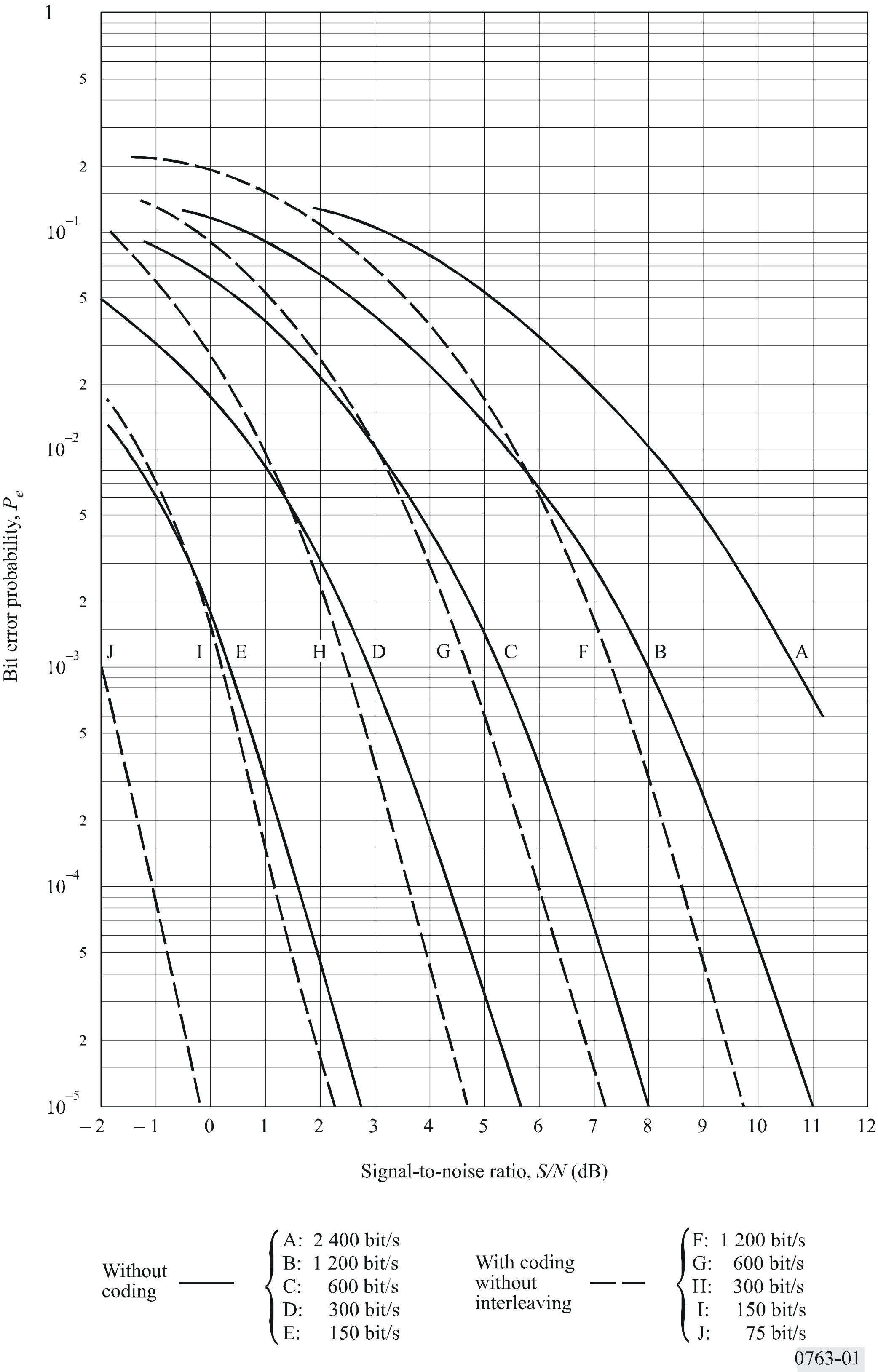
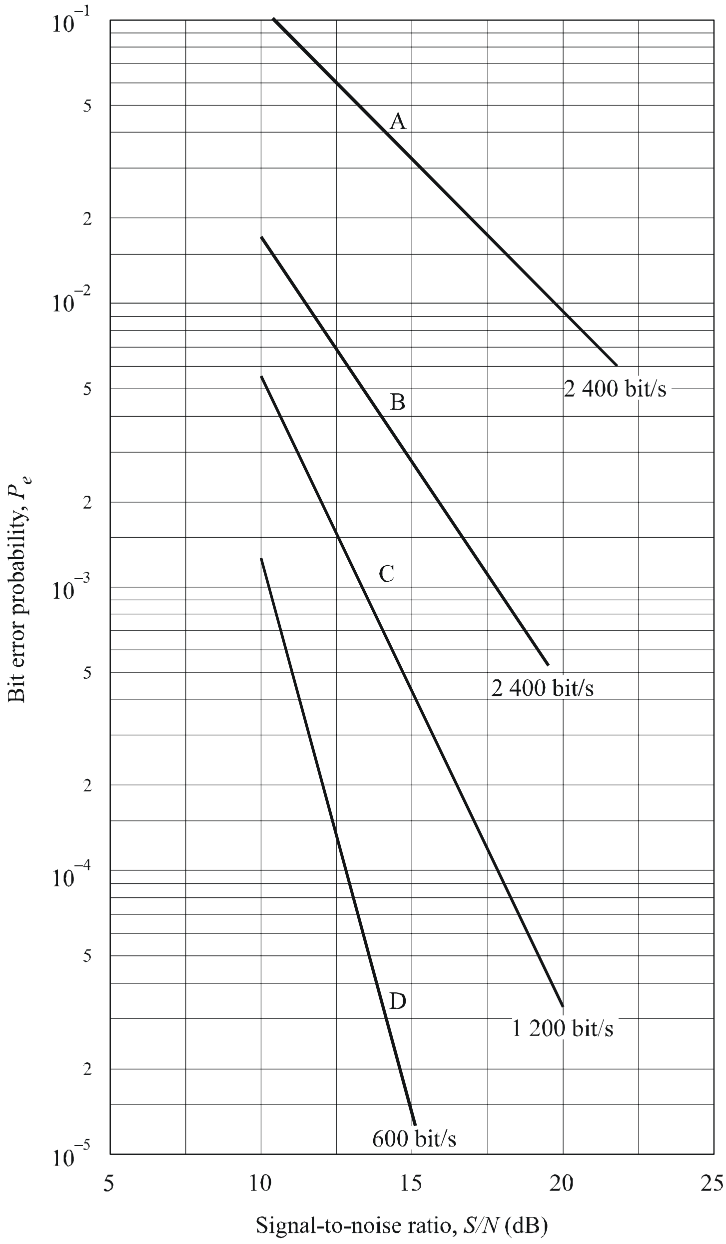


FIGURE 2

Bit error probability versus *S/N* for a selective fading channel using data rates of 600, 1 200 and 2 400 bit/s in the following cases



A: without diversity
B: out-of-band only
C: in- and out-of-band diversity
D: in- and out-of-band diversity with the use of error correcting codes and interleaving

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Annex 2

Data transmission at rates up to 3600 bit/s over HF circuits using a serial transmission modem

1 General

The modem permits data transmission in a 3 kHz HF channel. It receives and reconstitutes digital data at a rate of ≤ 3600 bit/s and generates an analogue AF signal within the 300-3300 Hz audio band.

It incorporates protection against multipaths, Doppler effect and fading.

2 Modem operating modes

There are three possible operating modes.

2.1 Semi-duplex forward error correction (FEC) mode

2.1.1 This mode uses an MPSK (M = 2, 4, 8) modulation at 2400 Bd, with a user bit rate of 75, 150, 300, 600, 1200, 2400 or 3600 bit/s (not all of the bit rates are available with all of the waveforms), and with frames of 256 modulated symbols (of which 128 are user symbols), i.e. 106.6 ms.

2.1.2 A data exchange comprises three phases, namely preamble, traffic and end of transmission:

FIGURE 3

Description of a communication in FEC mode

Preamble	Traffic (data)	End of transmission

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The preamble phase enables the called modem to detect the call and to receive the technical parameters (encoding, interleaving, data rate, modulation) that it needs for the rest of the transmission. The traffic phase contains the data to be transmitted. The end of transmission phase enables the called modem to detect an end of message word in order to terminate the link and return to traffic standby.

The end of transmission is effected when the calling modem transmits on-hook frames. These frames are similar to preamble frames, but include a bit containing the on-hook information.

2.1.3 The functions provided are as follows:

– *Emission:*
  – data encoding and interleaving;
  – framing and modulation;
  – transmission of AF signal.
– *Reception:*
  – reception of AF signal;

– detection of synchronization;
– demodulation of received signal;
– data de-interleaving and decoding.

2.2 Full-duplex FEC mode

This mode amounts to the same thing as two independent FEC-type semi-duplex links. In each direction a preamble followed by data and an end of message word are sent and recognized by the called modem. As in the semi-duplex FEC mode, this preamble specifies the technical parameters that are to follow.

2.3 Automatic repeat request (ARQ) mode

2.3.1 This mode uses an MPSK ($M = 2, 4, 8$) modulation at 2400 Bd, with a user bit rate of 600, 1200, 1800 or 2400 bit/s (not all of the bit rates are available with all of the waveforms), with frames of 256 modulated symbols (of which 128 are user symbols), i.e. 106.6 ms.

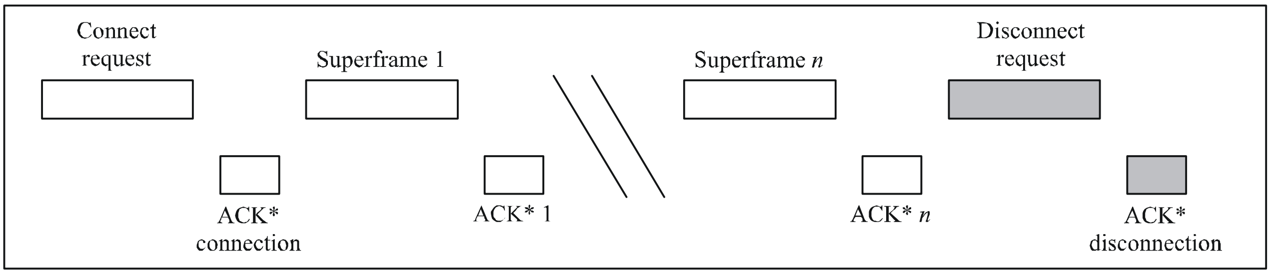
2.3.2 The ARQ mode is a data transmission mode involving selective repetition by block. The data for transmission are divided up into blocks corresponding to a modem frame. The calling modem sends a superframe of *N* blocks (*N* is nominally equal to 64, but may be lower than this during transmission of the last data) and waits for the called modem to acknowledge its receipt.

If any blocks have not been correctly received, they are re-transmitted in the following superframe, which is made up with new blocks.

The phases contained in this mode are call set-up (connection), data transmission and end of transmission (disconnection). In addition, the ARQ mode allows for momentary disconnection, caller/called party switching, flow control, and adaptive power, data rate and frequency control.

FIGURE 4

Description of a communication in ARQ mode



* ACK: acknowledgement

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The ARQ mode thus comprises two distinct phases, namely a transmission phase (transmission of a superframe at the calling end, and of an acknowledgement at the called end), and a reception phase (reception of an acknowledgement at the calling end, and of a superframe at the called end).

2.3.3 Adaptive control

2.3.3.1 The ARQ mode allows adaptive power, data rate and frequency control. Of these, only the adaptive data rate control is entirely managed by the modem. In the case of power control, the modem indicates to the system the adaptation to be effected and continues the transmission, while in the case of frequency control, the modem momentarily disconnects itself after indicating to the system the need to find a new frequency.

2.3.3.2 The adaptive power control procedure is based on statistical measurements of the link quality. Adaptive power increase is achieved very rapidly, while power decrease involves a large time constant.

2.3.3.3 Adaptive data rate control is effected on three of the data rates chosen from among the four that are available, namely 2 400, 1 800, 1 200 and 600 bit/s.

Adaptive increases in data rates are based on statistical measurements of the link quality, while decreases are based either on statistical measurement of the link quality, or on the non-reception of data or acknowledgements during the transmission.

2.3.3.4 If the adaptive data rate decrease control is not sufficient to continue the transmission, a request is made to the system to implement adaptive frequency control.

In order that a new frequency may be sought, the modem momentarily disconnects itself and stands by to resume the transmission, storing the data which have not yet been transmitted.

2.3.3.5 It is possible to set up the modem in ARQ mode in such a way that it does not implement adaptive data rate control. In this case, only the frequency and power control are effected.

2.3.4 The functions provided are as follows:

- *Send, at the calling end:*
  - data segmenting,
  - data encoding,
  - framing and modulation,
  - transmission of AF signal.
- *Send, at the called end:*
  - encoding of acknowledgements,
  - framing and modulation,
  - transmission of AF signal.
- *Receive, at the calling end:*
  - reception of AF signal,
  - detection of synchronization,
  - received signal demodulation,
  - decoding of acknowledgements.
- *Receive, at the called end:*
  - reception of AF signal,
  - detection of synchronization,
  - received signal demodulation,
  - decoding of data,
  - data reassembling.

3 Technical characteristics of the modem

3.1 Modulation

3.1.1 The modulation technique involves phase shift of a sub-carrier with a frequency of 1 800 Hz. The modulation rate is 2 400 Bd, with a minimum accuracy of 10^{-5}.

3.1.2 The clock stability associated with the generation of the 1 800 Hz is 10^{-5}.

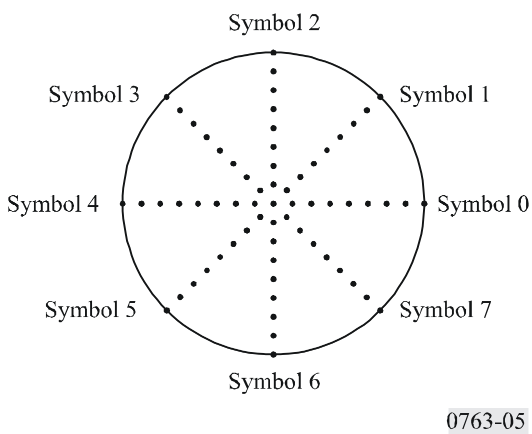
3.1.3 The phase shift of the modulated signal in relation to the unmodulated reference sub-carrier may take on one of the following values:

Symbol No.	Phase
0	0
1	$\pi/4$
2	$\pi/2$
3	$3\pi/4$
4	π
5	$5\pi/4$
6	$3\pi/2$
7	$7\pi/4$

Symbol number n is associated with the complex number exp ($jn\pi/4$).

FIGURE 5

Encoding of phase states



3.2 Transcoding

Transcoding is an operation in which a symbol to be transmitted is associated with a group of binary digits.

3.2.1 Data rate of 1 200 bit/s: 2-PSK

Transcoding is effected by associating a symbol with a binary digit according to the following rule:

Bit	Symbol
0	0
1	4

3.2.2 Data rate of 2 400 bit/s: 4-PSK

Transcoding is effected by associating a symbol with a set composed of two consecutive binary digits according to the following rule:

Dibit	Symbol
0 0	0
0 1	2
1 0	6
1 1	4

Oldest bit   Most recent bit

3.2.3 Data rate of 3 600 bit/s: 8-PSK

Transcoding is effected by associating a symbol with a set composed of three consecutive binary digits according to the following rule:

Tribit	Symbol
0 0 0	1
0 0 1	0
0 1 0	2
0 1 1	3
1 0 0	6
1 0 1	7
1 1 0	5
1 1 1	4

Oldest bit   Most recent bit

3.3 Frame structure

3.3.1 The symbols to be transmitted are structured in recurring frames of 106.6 ms in length. The number of binary digits transmitted per frame is 128 bits at 1 200 bit/s, 256 bits at 2 400 bit/s and 384 bits at 3 600 bit/s.

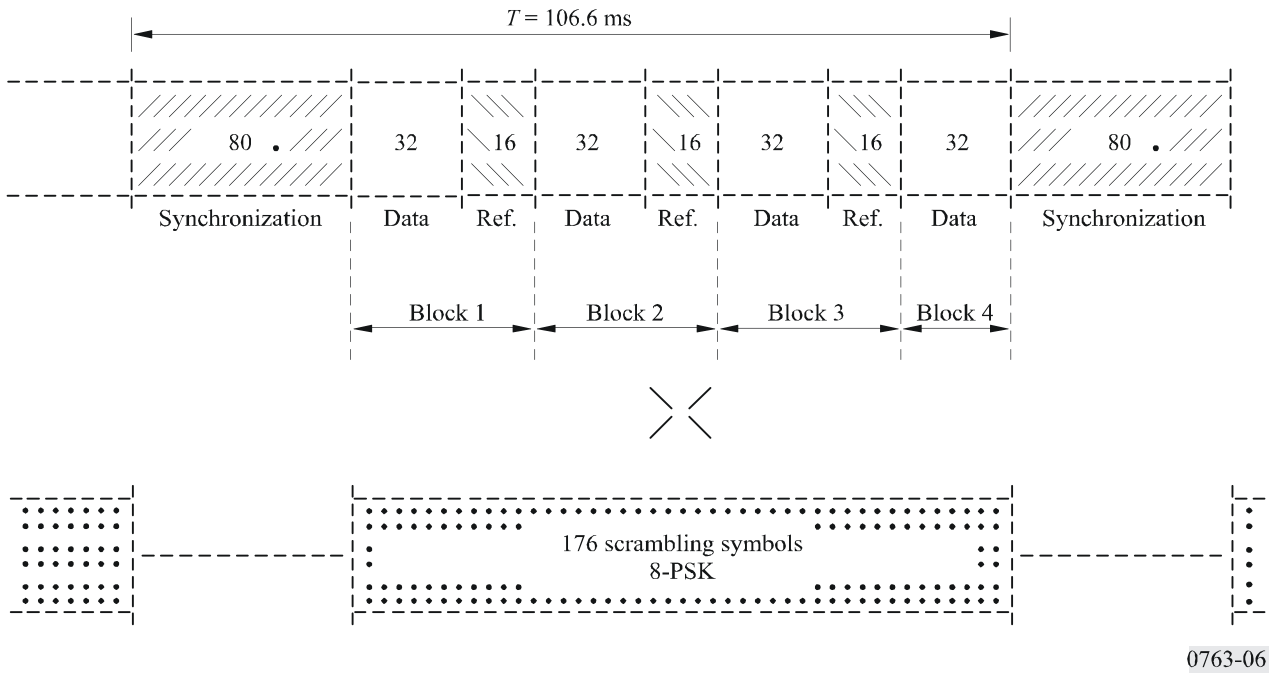
3.3.2 A frame is made up of 256 symbols, of which the breakdown is as follows: 80 symbols for synchronization, 48 reference symbols and 128 data symbols.

Figure 6 depicts the frame structure.

3.3.3 The synchronization sequence is transmitted in 2-PSK, at a modulation rate of 2 400 Bd. It is used by the modem for detecting the presence of the signal and for correcting the frequency shift resulting either from the Doppler effect or the difference between the transmit and receive carrier, bit synchronization and either the equalization time in the case of equalization by recursive filtering, or the HF channel evaluation in the case of detection by the maximum likelihood method.

FIGURE 6

Frame structure



3.3.4 The reference and data symbols are structured in four blocks, the first three of which comprise 32 data symbols followed by 16 reference symbols, with the last block comprising 32 data symbols. All of the reference symbols correspond to the symbol number 0.

These 176 reference and data symbols are scrambled by a scrambling sequence comprising 176 symbols which is repeated every 106.6 ms. This sequence is transmitted in 8 state phase modulation at the rate of 2 400 Bd. It is thus possible to create a frame with 8 phase states, whatever the data rate (1 200 bit/s, 2 400 bit/s or 3 600 bit/s).

The scrambling operation consists in adding modulo 8, the symbol number associated with the data to the symbol number associated with the scrambling, which amounts to complex multiplication of the data symbol by the scrambling symbol.

3.4 Error correction coding, interleaving

The use of error correction coding in conjunction with adequate interleaving can considerably improve the BER.

On the basis of the three base modes without redundancy, namely

– 3 600 bit/s 8-PSK,
– 2 400 bit/s 4-PSK,
– 1 200 bit/s 2-PSK,

the coding permits the introduction of various redundancy possibilities.

3.4.1 FEC mode

This involves the use of convolutional coding in combination with interleaving which is also convolutional. The convolutional code used is redundant code 2 and constraint length K = 7, associated with the characteristics polynomial 171,133 (octal representation).

Redundancies lower than 2 are obtained by puncturing, while redundancies higher than 2 are obtained by repetition.

Among the various possibilities, we would mention the following:

Data rate with coding (bit/s)	Waveform	Redundancy	Method for obtaining this code rate
2 400	8-PSK	3/2	Conversion of data rate 1/2 to data rate 2/3
1 200	4-PSK	2	Unmodified code at data rate 1/2
600	2-PSK	2	Unmodified code at data rate 1/2
300	2-PSK	4	Code at data rate 1/2 repeated 2 times
150	2-PSK	8	Code at data rate 1/2 repeated 4 times
75	2-PSK	16	Code at data rate 1/2 repeated 8 times

3.4.2 ARQ mode

A Reed-Solomon (RS) coding is used, and there is no interleaving.

Data rate with coding (bit/s)	Waveform	Redundancy	Coding (symbols of 8 bits)
2 400	8-PSK	3/2	RS (48,32)
1 800	4-PSK	4/3	RS (32,24)
1 200	4-PSK	2	RS (32,16)
600	4-PSK	4	RS (32,8)

3.5 Spectrum of the modulated signal

The spectrum of the modulated signal after filtering and 1 800 Hz transposition is shown in Fig. 7. The total bandwidth is equal to 3 000 Hz.

3.6 Frequency error tolerance between transmission and reception HF carriers

The modem must be able to tolerate a shift of ±75 Hz between the transmission and reception HF carriers (transmitter/receiver frequency error and Doppler shift included) and a frequency variation rate of at most 3.5 Hz/s.

4 Interfaces with other equipment

4.1 Modem interface with the data terminal

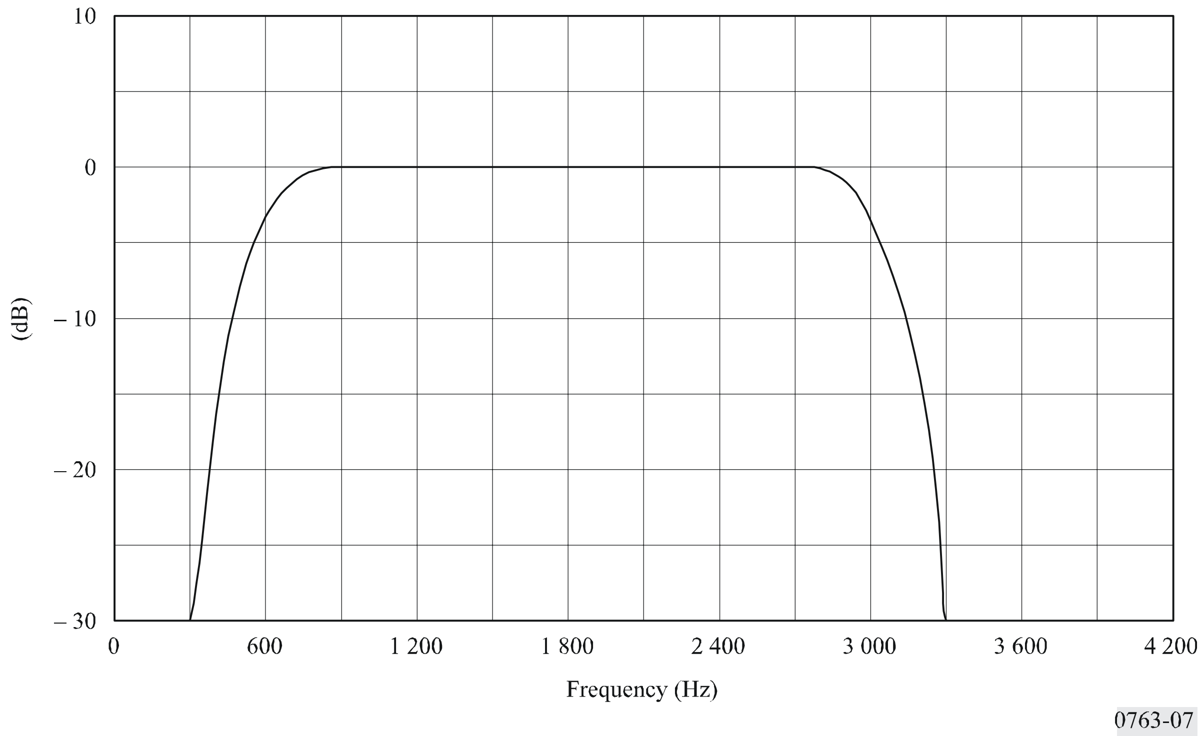
This satisfies ITU-T Recommendation V.24, the electrical characteristics of the interface being in conformity with ITU-T Recommendation V.11 (RS 422).

4.2 Modem interface with the transmitter and the receiver

The input and output circuits of the modem are of the balanced to earth type, having an impedance of 600 Ω at 0 dBm.

FIGURE 7

Spectrum of the modulated signal



4.3 Quality of performance of associated transmitters and receivers

To obtain optimal performance, the following characteristics for transmitters and receivers are recommended:

4.3.1 They must have a passband such that between 300 Hz and 3 300 Hz variations of transmission loss are at most ± 2 dB.

NOTE 1 – The operation of a serial modem with a system bandwidth of 300 to 3 000 Hz is possible with reduced performances. Further study would be needed to design a serial modem with a sub-carrier of 1 650 Hz, operating with reduced bandwidth systems.

4.3.2 The group delay must not vary by more than 0.5 ms over 80% of this passband.

4.3.3 The accuracy of the transmitter and receiver pilot frequencies must be at least 10^{-6}.

4.3.4 The time constant of the automatic gain control (AGC) circuit must be less than 10 ms for de-sensitization and less than 25 ms for re-sensitization.

Annex 3

Transmission systems using PSK

1 Introduction

In HF channels information at bit rates of over 200 bit/s is normally transmitted using multistate methods and complex signals. This generally involves a combination of frequency-shifted orthogonal subcarriers with 2-PSK. With the latter technique a bit rate twice that obtainable with FSK can be achieved in the same frequency band and the redundancy can be used to increase noise immunity. Apart from multi-frequency PSK, practical interest attaches to a more general type of modulation – generalized PSK, in which the information to be transmitted is contained not in the differences between the instantaneous phases of the sine-wave signals but in the difference between the phase spectra of complex orthogonal signals. The amplitude spectra of such signals coincide and may be matched with the channel frequency characteristic (or the interference spectrum) without violating the conditions of mutual orthogonality. On this basis it is possible to consider the construction of adaptive modems with a higher noise immunity or traffic capacity.

The practical application of generalized PSK has been held up in the past by the well-known difficulties involved in the synthesis and processing of complex signals. The basic problems have now been solved, thanks to the theory of synthesis which has been developed, and the availability of microelectronic modules with a high degree of integration, which has removed the obstacle of technical circuit complexity. This Annex sets forth the main principles governing the design of modems with generalized PSK, describes a variant which has been developed and gives a number of test results.

2 Theoretical questions

2.1 Selection of signals

As was pointed out by Shannon, in order to achieve a transmission rate equal to the communication capacity in channels with a frequency characteristic of $Y(\omega)$ and a Gaussian noise of $N(\omega)$, use has to be made of signals characterized by a steady state Gaussian process with a power of P and a power spectrum of:

$$F(\omega) = \begin{cases} B - \dfrac{Y(\omega)}{N(\omega)} & \text{for} \quad \omega \in \Omega \\ 0 & \text{for} \quad \omega \notin \Omega \end{cases} \tag{1}$$

where the integration range Ω is determined from the condition $F(\omega) \geq 0$ and the constant B depends on the power of the signals. Since in practice there are always standards governing the permissible limits for the delay in information being transmitted, the maximum signal duration and the number of signals have to be limited. Under these conditions finite-dimensional combinations of determined orthogonal signals, the squares of whose spectral density moduli coincide with $F(\omega)$, may be considered as being close to optimum. However from (1) it follows that $F(\omega) = 0$ at all frequencies where $B < Y(\omega)/N(\omega)$, i.e. mutual orthogonality has to be preserved when individual portions of the spectrum are rejected. The multi-frequency signals used in existing modems do not possess this property. Moreover, their orthogonal spectrum shape is optimum only for channels with a flat frequency characteristic and white noise type interference. Calculations show that failure to meet these conditions may result in information transmission rate losses of up to 40% of the channel's communication capacity.

Another criterion for assessing the optimum nature of orthogonal signal combinations is the requirement as regards the shape of their autocorrelation function. For example, to ensure stability in the operation of a synchronization system, the main lobe of this function has to be narrow enough and the side lobes must not exceed a given level. In this case mutual orthogonality must be assured at a given signal amplitude spectrum which does not necessarily satisfy condition (1).

In view of the above, in order to achieve generalized PSK, a special class of signals based on the use of complex systems of functions with double orthogonality was developed. Their spectral densities may be represented as follows:

$$S_k(\omega) = |S(\omega)| \, e^{j[K\,\psi(\omega) + \alpha(\omega)]} \tag{2}$$

where:

$$|S(\omega)|^2 = A \left| \frac{d\psi(\omega)}{d\omega} \right|$$

where:

- A: constant factor
- $\alpha(\omega)$: arbitrary bounded function.

For a given amplitude spectrum, it is therefore possible to define the phase spectrum of the signals and therefore their spectral density. Further synthesis involves finding samples of spectral signal densities with different serial numbers and transforming them using a Fast Fourier Transform (FFT) into time samples. Synthesis of the signals may be combined with coding of the signals in the time domain using the Reed-Solomon code; for this purpose a number of zero samples has to be added beforehand to the spectral density samples and only then can the FFT be performed. It should be noted that this type of mixed coding (orthogonal in the frequency domain and Reed-Solomon code in the time domain) is most effective for HF channels.

2.2 Selection of a processing algorithm

In the case of multi-state methods for the transmission of information, it is best to process the signals to be received using the optimum algorithm for reception "as a whole". The simplest way of implementing such an algorithm is to use component demodulators; for this purpose the following conditions have to be met:

– the multi-state signals have to be component-type, i.e. they have to be made up of the sum of the elementary signals;

– each elementary signal must contain information about the corresponding element of a code word $b_{i,k}$;

– the interference affecting the elementary signals must be mutually independent.

In this case the decision rule is as follows:

$$\max\left[L_i = \sum_{k=1}^{N} e_{i,k} y_k\right] \tag{3}$$

where:

$e_{i,k}$: sign coefficient which takes the value: +1 when $b_{i,k} = 1$, and –1 when $b_{i,k} = 0$

$$y_k = \ell_n \frac{W(Z_{k/1})}{W(Z_{k/0})}$$

where:

Z_k: complex input signal (see Fig. 1)

$W(Z_{k/1})$: probability of Z_k being 1

$W(Z_{k/0})$: probability of Z_k being 0.

Optimality in this case is determined by the extent to which the signals used meet the conditions listed above. The first two involve the possibility of using a component demodulator. For these conditions to be met, it is sufficient for each spectral density sample (or its components) to contain information about the sign of the corresponding binary symbol. The condition stipulating the mutual independence of interference can be reduced to a condition stipulating the independence of the projections of the received signal vector onto the system on the basis of Fourier transform functions. This condition is fulfilled in the case of independent fades in individual frequency bands, invariance of the basic functions to time shifts and interference with a flat power spectrum. In practice it is impossible for the various requirements listed above to be met fully, so that the noise immunity of the component demodulator will be lower than the potential noise immunity although much higher than in the case of separate signal element reception.

The block diagram of the receiving section of the modem which implements decision rule (3) consists of the following units (see Fig. 8): a unit for calculating the logarithm of the likelihood ratio y_k; a unit for calculating the linear forms L_i; a decision element determining the number of the linear form with the maximum value; a digital converter which compares with each number its own combination of binary symbols; and this provides an assessment of the information sequence transmitted.

3 Description of the system

A block diagram of the system is shown in Fig. 9. It consists of the following elements: the user terminals; the signal conversion unit (modem) located either in the immediate vicinity of the terminals or in a separate communication control unit; the receiving and transmitting SSB equipment and the corresponding antennas. When the modem is installed in the control unit, communication with the terminal is established via tone frequency channels.

When it is set up in the immediate vicinity of the terminal, it can be connected up by d.c. circuits.

FIGURE 8

Receiving section of modem

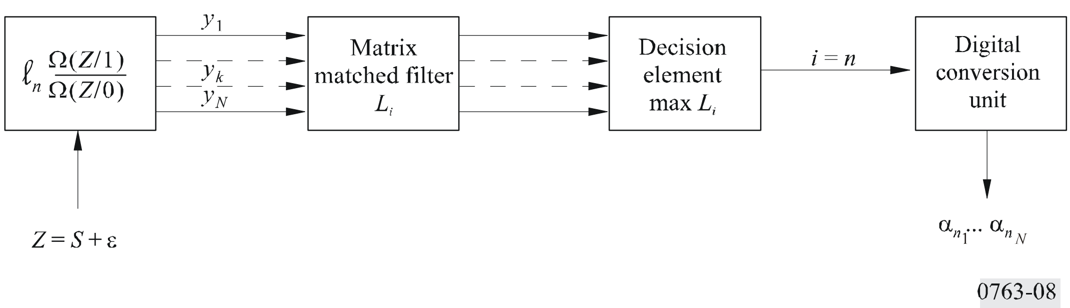
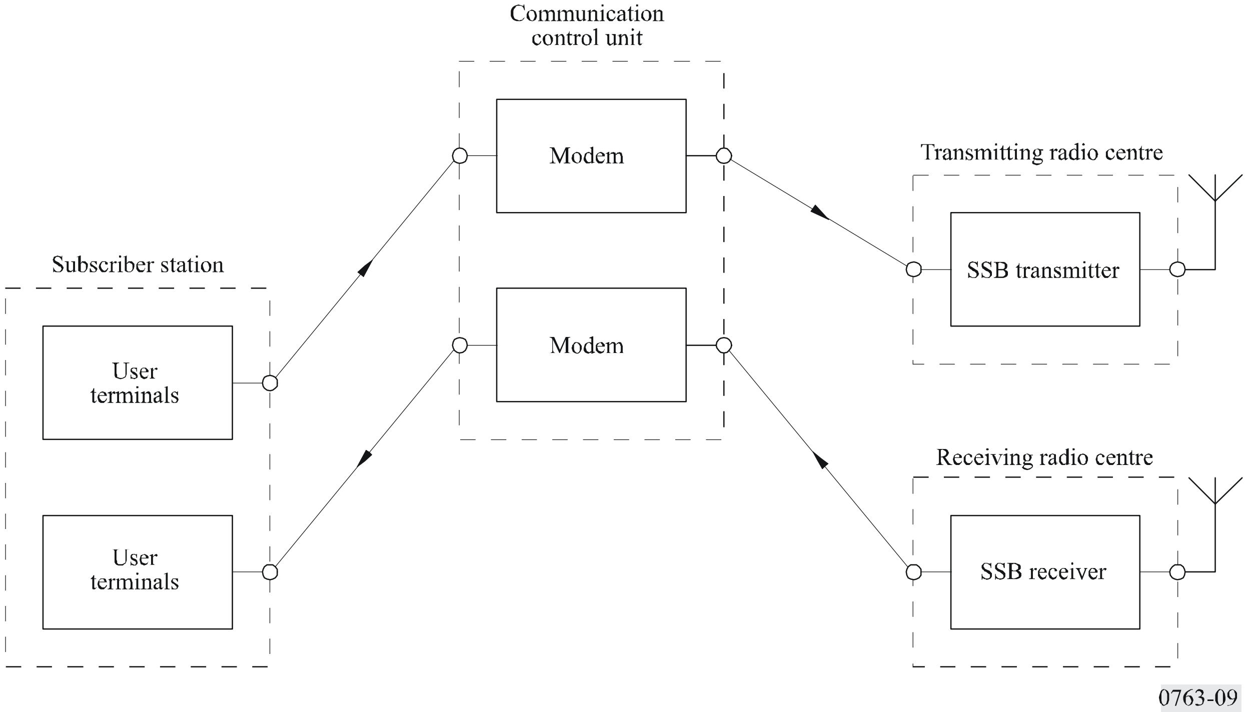


FIGURE 9

Block diagram of system



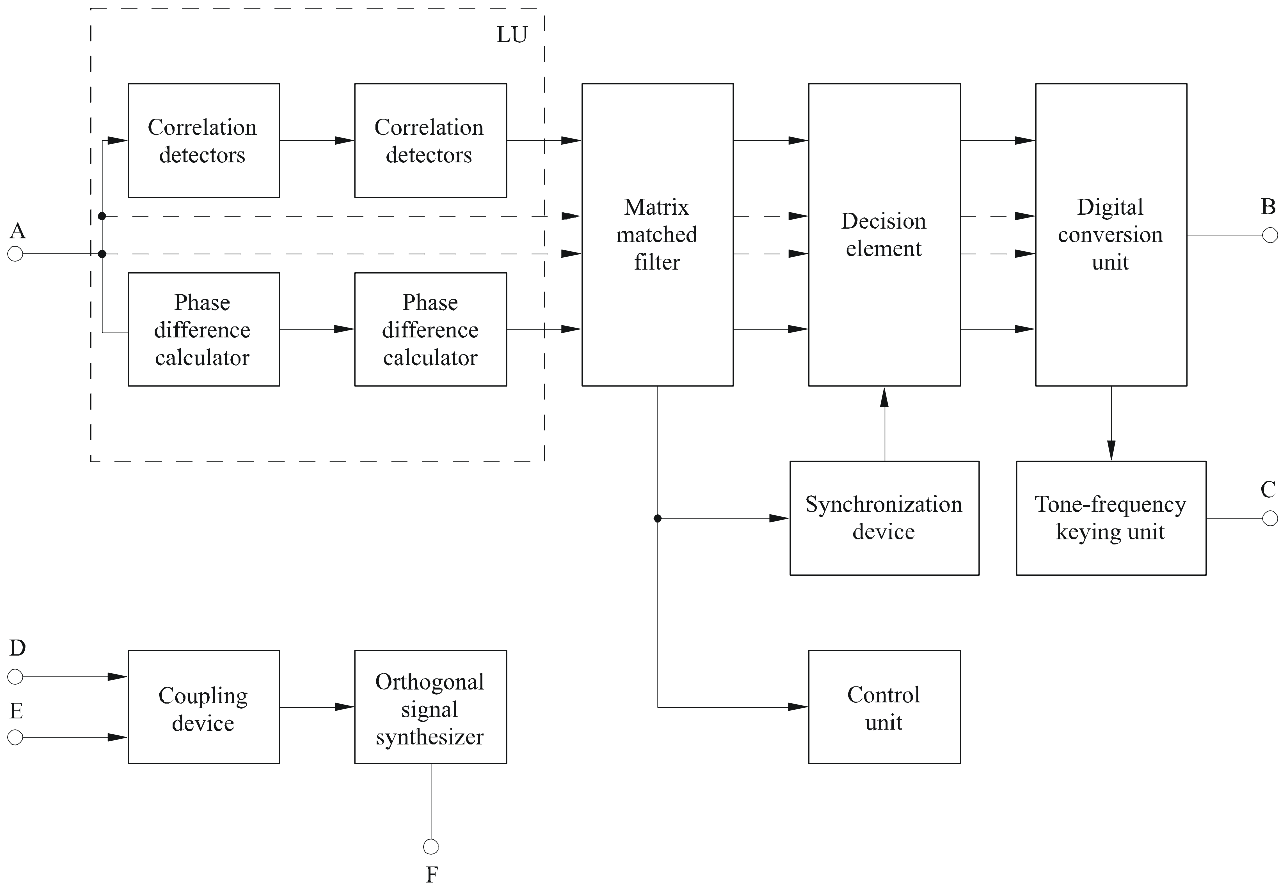
A block diagram of the modem embodying the principles considered above is shown in Fig. 10. The modem is designed to transmit digital information at 600-1 200 bit/s. For lower bit rates, an additional codec has to be used. A bit rate of 2 400 bit/s is obtained by increasing the number of signals used and switching over to separate signal element reception. The modem transmitter consists of a coupling device (CD) and an orthogonal signal synthesizer (OSS).

The CD is designed to match the modem with the user's terminal via the tone frequency channels or the d.c. circuits and to control the synthesizer. It includes a tone frequency amplifier-rectifier, a regenerator, and a logic control circuit.

The OSS shapes the analogue signals and amplifies them to the requisite level. It consists of a coding unit, a ROM, a digital to analogue converter (DAC), an low frequency filter and a power amplifier. A specific feature of the operation of the OSS is that time samples of all the signals to be used for the transmission of information are already entered in its ROM. These samples were calculated beforehand on a computer in accordance with the rules laid down in the previous section.

FIGURE 10

Block diagram of modem



A: modem receiver input
B: modem receiver tone frequency output
C: d.c. modem receiver output
D: tone frequency input of modem transmitter
E: d.c. input of modem transmitter
F: modem transmitter output

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In an initial stage, in order to verify the basic principles used, a set of 16 biorthogonal signals was synthesized; these had a flat amplitude spectrum in the range 1.1-2.42 kHz and an effective band of 0.66-2.86 kHz. Their spectral densities were represented by means of four complex samples, each of which could provide information on the signs of two binary symbols. To transpose the spectrum to these samples, two zero samples were added and after a Fourier transform an additional multiplication by a complex component was carried out.

The time samples of the signals calculated in this way were entered in an 8 bit grid in the ROM and, after they had been read at a timing frequency of 8.8 kHz, it was possible to obtain at the output of the DAC unit analogue signals with a duration of 3.33 ms and an orthogonality interval of 2.27 ms.

The sequence of operations in the modem transmitter is as follows: the binary information signals from the terminal are regenerated, combined to form 4-bit code words and are then fed to the input of the relative coding circuit which controls the selection from the ROM of one of the 16 forms of the signal. From the output of the ROM the samples are converted by the DAC unit into an analogue signal which after being amplified is fed along the tone frequency channel to the input of the SSB transmitter.

As shown in Fig. 8 the receiving section of the modem consists of the following elements: a unit which calculates the logarithms of the likelihood ratio (correlation detectors (CDT); a phase difference calculator (PDC)); a matrix matched filter (MF) which calculates all the linear forms L_i; a decision element (DE) which determines the number of the maximum form; and a digital conversion unit (DCU). It also contains a synchronization device (SD) and a control unit (CU). Provision is made for a single or a dual operating mode with space or polarization diversity.

The transformation of the analogue signals into spectral density samples is carried out by the correlation detectors which calculate the in-phase and the quadrature component of each sample. The initial uncertainty relating to the phase of the channel is then eliminated using the PDC and the phase spectrum of the signal received is calculated. The matched filter is a matrix adder and each of its columns is adjusted for an appropriate sample selection using inverting amplifiers. The DE looks for the column with the maximum output voltage and, using the DCU, transmits the corresponding 4 element sequence of binary symbols which is fed to the input terminal either directly or via the tone-frequency keying unit (TKU).

The CU works on the principle that the voltages at the output buses of the matched filter coincide precisely to within a constant factor with the distribution of *a priori* probabilities. It is clear that the performance of the channel will be better the “steeper” this distribution is since in the ideal case the voltage must appear only on one of the output buses of the matched filter. The difference between the maximum voltage and the voltage whose level is closest to it at the other bus may be used to evaluate the quality of the channel in the information transmission process.

4 Experimental investigations

The laboratory tests on the modem were carried out using a modelling test bench which included the following elements: the SSB receiver; a two-ray channel simulator; a noise generator; and a digital counter to calculate the number of errors. A pseudo-random sequence (PRS) from a generator built into the modem was used as a test combination. Three modes of operation were analysed: a channel with constant parameters and white noise; a single-ray channel with Rayleigh fading; and a two-ray channel with a difference in the ray propagation time of 1 ms, identical ray amplitudes and Rayleigh fading. The results of the tests are shown in Figs. 11 and 12. By way of comparison, Fig. 11 shows curves for the noise immunity of a multi-frequency modem as described in Annex 1 for the same transmission speed. As the curves show, the modem studied has a better noise immunity. Comparison of curves A and B in Fig. 12a) show that the modem has a higher noise immunity in the two-ray channel than in the one-ray channel. The reasons for this is that in the case of flat fades, decision rule (3) is no longer optimum. In a two-ray channel a predominant role is played by frequency selective fades which the modem can combat more effectively. The dotted line in the figure shows the theoretical curve for the noise immunity of an optimum non-coherent separate signal element reception using binary PSK in the case of Rayleigh fading.

Link tests on the modem were carried out on 3 600 km and 4 300 km latitudinal paths. Use was made of a 15 kW SSB transmitter, rhombic transmitting antennas and fishbone receiving antennas (dual reception). Tests on the first path were carried out during the day and during the night on one frequency. On the second path two frequencies were used. The information bit rate was 1 200 bit/s.

On the basis of 5 min measurements, integral curves were plotted showing the error rate distribution; these are shown in Fig. 12b).

5 Conclusions

The use of generalized PSK in combination with reception "as a whole" opens up additional possibilities for increasing noise immunity in the transmission of digital information. A modem developed to serve as a practical example of how the generalized PSK method may be implemented uses signals with a flat spectrum and from this point of view is similar to the modems described in Annex 1. Tests have shown that for links over 3 000-4 000 km it guarantees a bit rate of 1 200 bit/s with an error rate not exceeding 1×10^{-4} to 1×10^{-3} for 95 to 98% of the time.

FIGURE 11

Noise immunity of modem

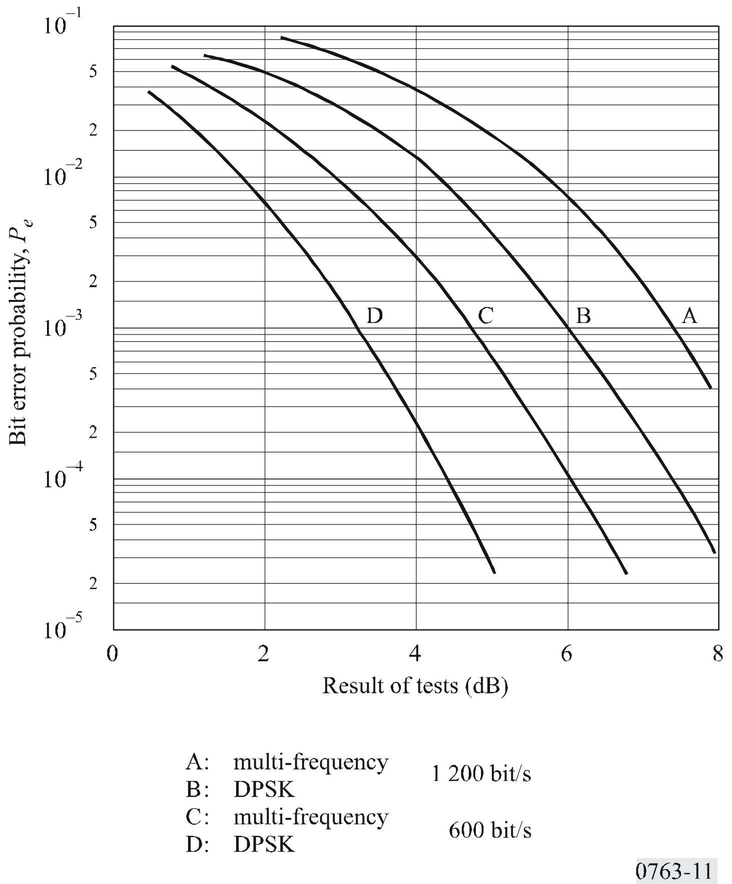
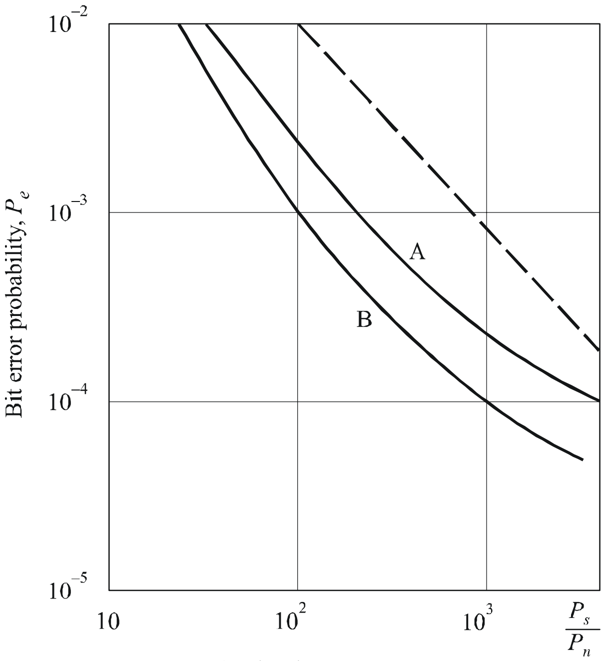


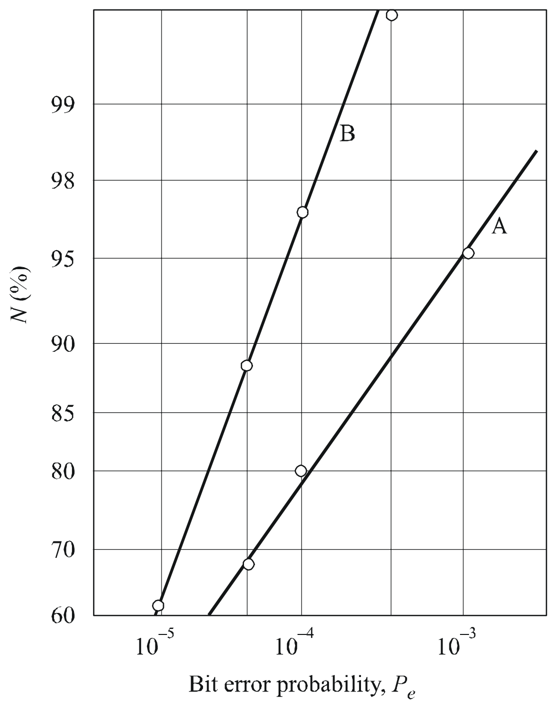
FIGURE 12

Noise immunity of modem in fading channels



a) Simulator tests

A: 1 ray
B: 2 ray
Transmission speed: 1 200 bit/s



b) Communication link tests

A: 3 600 km
B: 4 300 km
Transmission speed: 1 200 bit/s

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Annex 4

Mode/polarization diversity in high-frequency radio data systems

1 Introduction

The amplitude of a received HF radio signal fluctuates when its direction of polarization changes with respect to the receiving antenna with minima occurring when the polarization is orthogonal and maxima when the polarization is parallel. Fading due to polarization changes has been confirmed by experiments which have found that a minimum received signal level on one antenna element often coincides with a maximum signal level on an orthogonal element. This effect may be exploited by using a system of orthogonal antenna elements to improve system performance.

Many serial-tone HF modems incorporate adaptive equalization techniques such as those described in Annex 2. Some modems use a waveform in which a preamble is inserted periodically in the data stream. The preamble, which consists of known symbols, permits the instantaneous impulse response of the channel to be estimated. An adaptive equalizer can then use the estimated impulse response to combine energy from different paths having different delays. The impulse response is kept current through a least mean-squares update procedure to update the adaptive equalizer.

After equalization, the existence of several distinct propagation modes can be advantageous since they are unlikely to suffer from simultaneous fades thereby increasing the probability that some of the transmitted energy will be received. This phenomenon, known as mode diversity, can be exploited so long as the transmitted energy arriving at the receiver is sufficient to overcome the noise. Mode diversity gain can be best utilized if the path delay difference is large enough to avoid flat fading. By using the orthogonal antenna elements an artificial multipath of some fixed value can be created at the input of the demodulator. In this manner, polarization diversity gain can be achieved by taking advantage of the modem's ability to cope with inter symbol interference and improve performance through mode diversity.

Two different techniques have been considered. The first so called transmit diversity, uses two orthogonal antennas each driven by a separate but phase and frequency locked transmitter, with the baseband input to one of the transmitter delayed, and communicating with a receiver having a single antenna. The second, receive diversity, uses a single transmitter and antenna but two phase and frequency locked receivers connected to orthogonally polarized antennas. The receiver outputs were connected to a diversity combiner, again with one path delayed at baseband, which produced the input to the modem. The receiver outputs connected to a diversity combiner whose function was to combine the two signals to form the input signal to the modem. This simple combiner allows for receive diversity without modification to the modem. HF receivers usually employ an AGC to accommodate the wide dynamic range of a signal, so as to maintain an output close to some set level. When the input signal level to the receiver is reduced during a fade, the receiver gain is increased by the AGC action. The AGC voltage is therefore a convenient measure of instantaneous *S*/*N*. The design of the combiner should emphasize the component with better *S*/*N* at the expense of the component with the poor *S*/*N*. For this reason the AGC voltages of the receivers are utilized by the diversity combiner to determine the proportion of the two signals making up the sum. The resulting signal was then applied to the modem input.

For the system described in Annex 2, where the equalizer capability extends over 5 ms, a baseband delay of 2.7 ms has been found to be optimum. It has been found that the best results are obtained when the delayed path was the weaker path. This was due to the particular synchronization techniques used in the modem. For this reason, the procedure of using signal delay with the vertical antennas ensured the stronger signal preceded the weaker signal in both techniques.

2 Conclusions

This kind of diversity can significantly improve the performance of HF radio data systems. Transmit diversity can reduce the error rate by up to four orders of magnitude while receive diversity can improve the error rate by up to three orders of magnitude. The improvement offered by polarization diversity may be assessed by considering the amount of additional transmit power required to improve the performance of a system without diversity to the level achieved with diversity. For a modem incorporating adaptive equalization, the use of transmit diversity is equivalent to a transmitter power increase of about 6-8 dB while the simple receive diversity is equivalent to a power increase of 3-4 dB. For a system employing transmitter diversity two 100 W transmitters could replace a 1 kW transmitter if 7 dB gain is achieved. This reduction in transmitter power coupled with the fact that polarization diversity can be implemented at either the transmit or receive end of a link without modification to existing modems, could represent a significant cost reduction. The type of diversity employed in a particular application will be dependent on the type of link involved. That is, the base station would likely employ diversity, while a remote station would not. Transmit and receive diversity in particular are useful when the performance of data communication links to mobile platforms or remote sites can be enhanced with additional antennas, receivers and transmitters at the base station location.

Annex 5

Data transmission at rates up to 4 800 bit/s over HF circuits using a serial PSK or quadrature amplitude modulation (QAM) transmission modem

1 General

This modem permits data transmission with information rates of up to 4 800 bit/s using 16-QAM within a 300-2 700 Hz bandwidth. The modulation method is switched, according to link quality, to QPSK at 2 400 bit/s or to BPSK at 1 200 bit/s.

2 Features

– Information rates of up to 4 800 bit/s are available.
– The information rate is switched to 2 400 bit/s (with QPSK) or 1 200 bit/s (with BPSK) according to link quality.
– The transmission bandwidth is within 300-2 700 Hz which permits 3 kHz channel separation.
– The protocol includes a synchronizing sequence of 28 symbols related to each data frame at 112 symbols so that the raw bit rates for transmission are 6 kbit/s, 3 kbit/s and 1.5 kbit/s.
– Switching of the bit rate with modulation mode is smoothly achieved only by mapping switch without changing the signalling rate.
– A bidirectional decision feedback equalizer (DFE) is used.

3 Specification

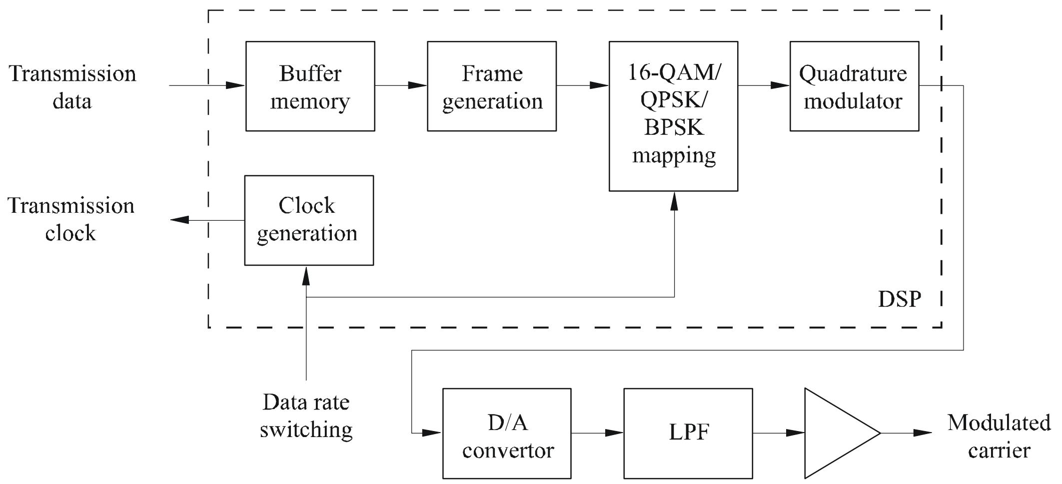
Modulation mode	16-QAM	QPSK	BPSK
Carrier bit rate (kbit/s)	6	3	1.5
User bit rate (kbit/s)	4.8	2.4	1.2
Signalling rate (kBd)	1.5		
Frame length	140 symbols (93.3 ms)		
Synchronization sequence	28 symbols		
Data length	112 symbols		
Equalization	Bidirectional DFE		

4 Block diagram of signal processing

Figures 13a and 13b show block diagrams of modulator and demodulator, respectively.

FIGURE 13a

Block diagram of modulator



DSP: digital signal processor
LPF: low pass filter

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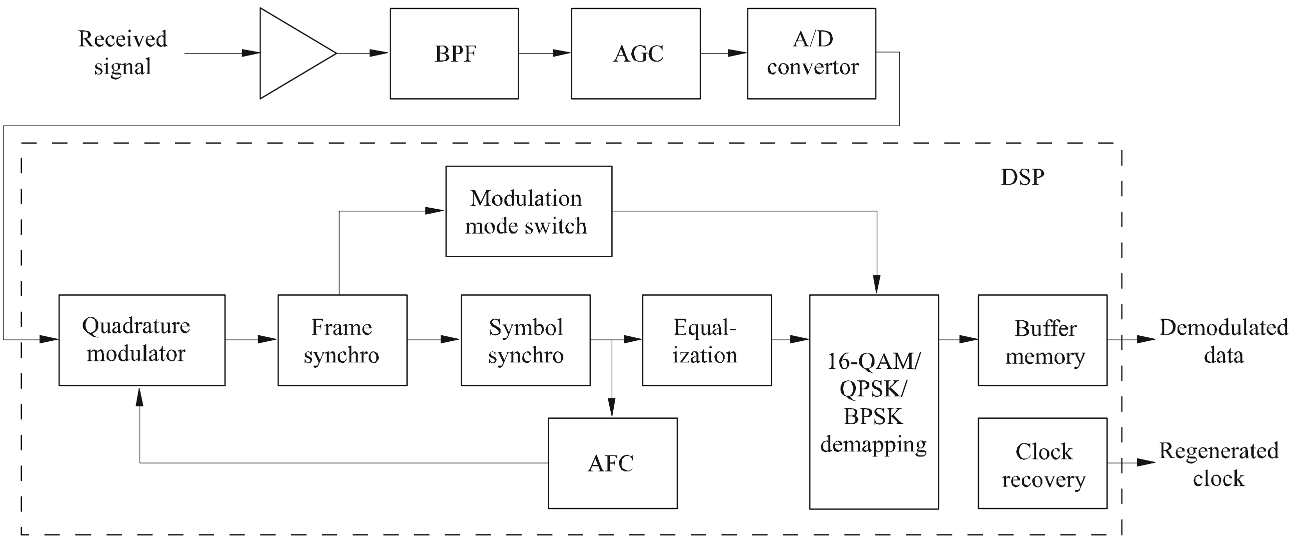
FIGURE 13b

Block diagram of demodulator



AFC: automatic frequency control
BPF: band-pass filter

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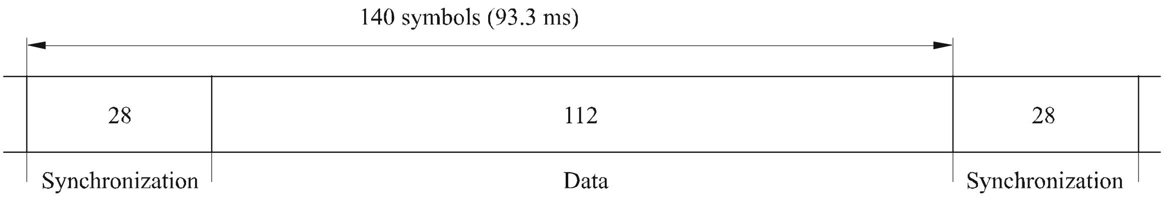
5 Frame structure

The symbols to be transmitted are structured in recurring frames of 93.3 ms in length as shown in Fig. 14.

FIGURE 14

Frame structure



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6 Coding rule and constellation diagram of 16-QAM

Table 2 and Fig. 15 show the coding rule for 16-QAM and the constellation diagram of 16-QAM, respectively.

TABLE 2

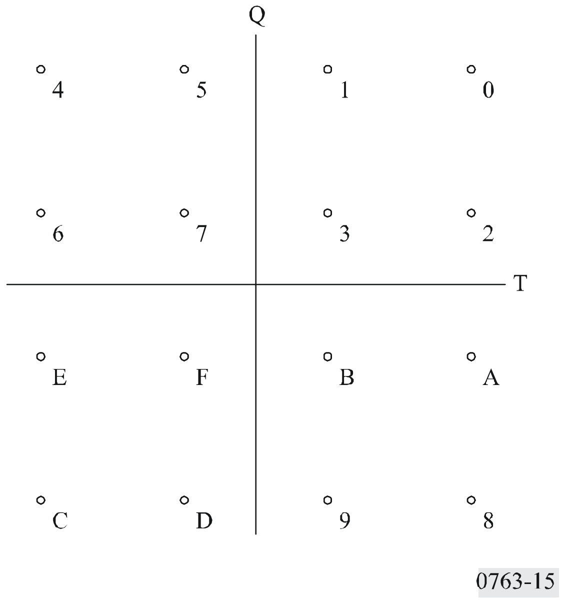
Coding rule of 16-QAM

Tetrabit	Symbol
0 0 0 0	0
0 0 0 1	1
0 0 1 0	2
0 0 1 1	3
0 1 0 0	4
0 1 0 1	5
0 1 1 0	6
0 1 1 1	7
1 0 0 0	8
1 0 0 1	9
1 0 1 0	A
1 0 1 1	B
1 1 0 0	C
1 1 0 1	D
1 1 1 0	E
1 1 1 1	F

Oldest bit   Most recent bit

FIGURE 15

Constellation diagram of 16-QAM



7 Test data

In the test described below, the DFE equalizer used 14 feedforward taps and six feedback taps with the capability of equalizing over a maximum delay of five symbols. Figure 16 shows the non-fading tests in Gaussian noise. The fading tests were conducted in accordance with Recommendation ITU-R F.520 with equal path gains and path delay differences of 0.5-3 ms, and fading rate of 0.5 Hz. Figures 17-19 show the bit error test results in the fading environment.

FIGURE 16

BER versus noise spectral density for a no-fading channel with Gaussian noise

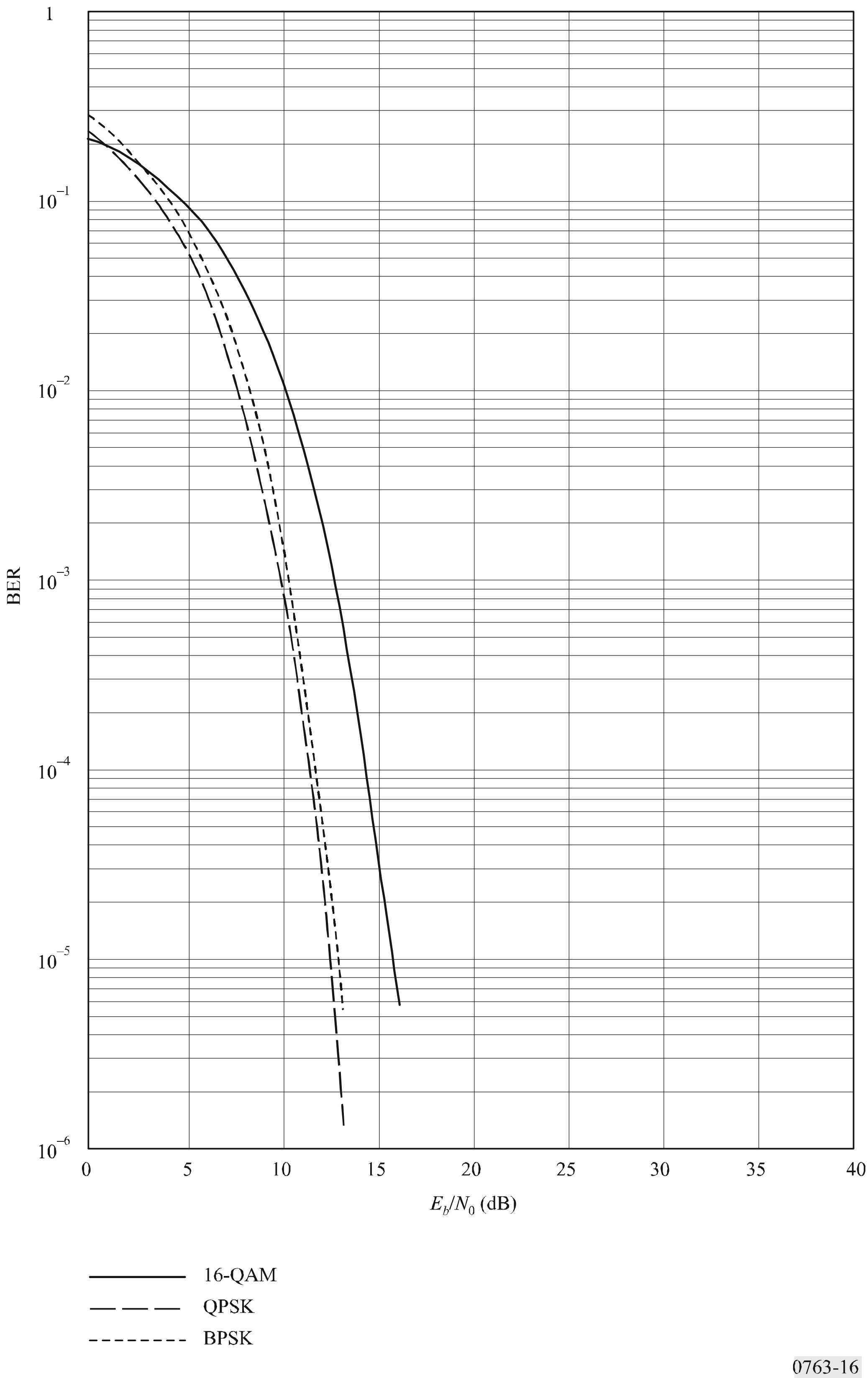


FIGURE 17

16-QAM BER versus noise spectral density for a fading channel

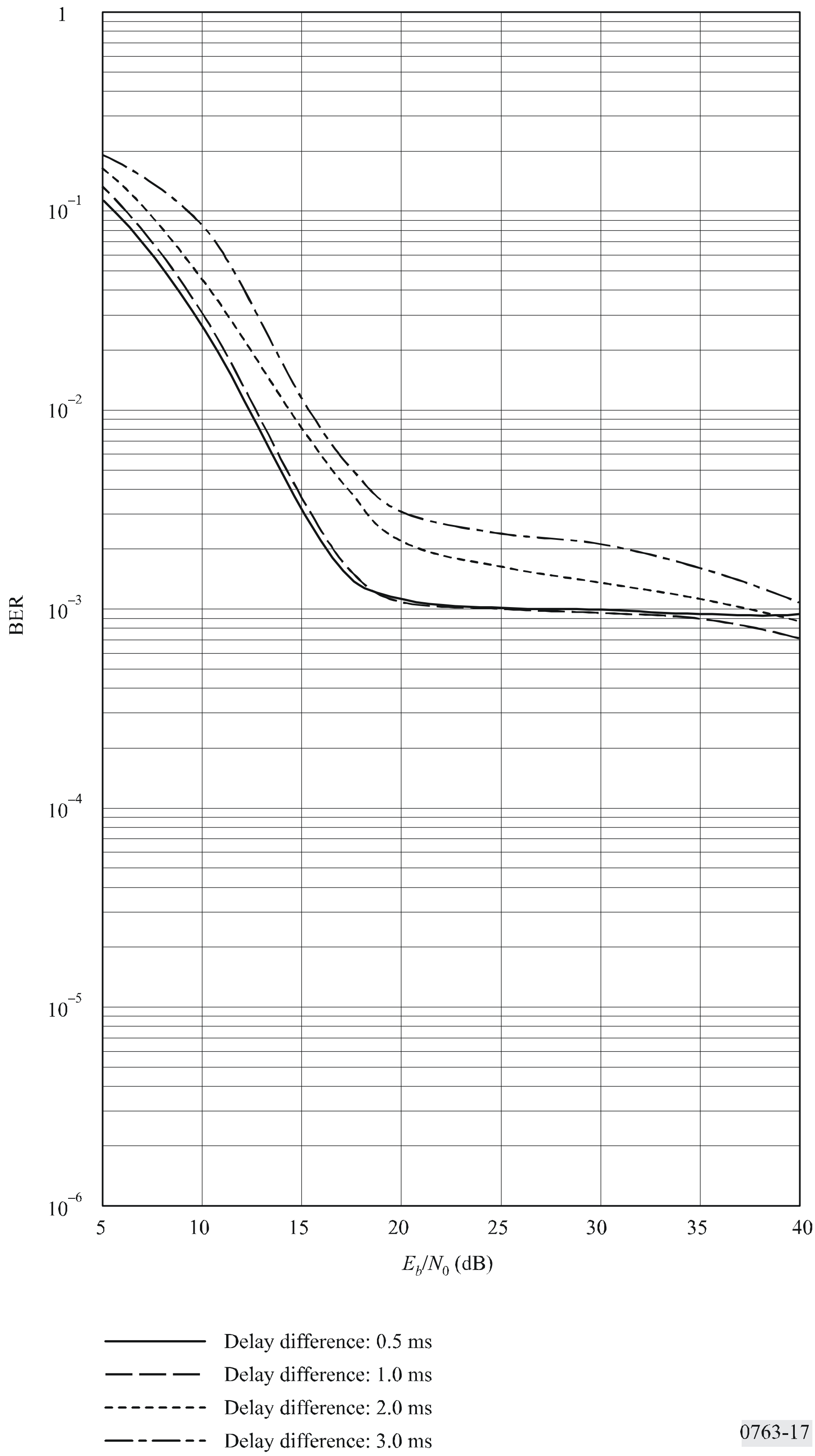


FIGURE 18

QPSK BER versus noise spectral density for a fading channel

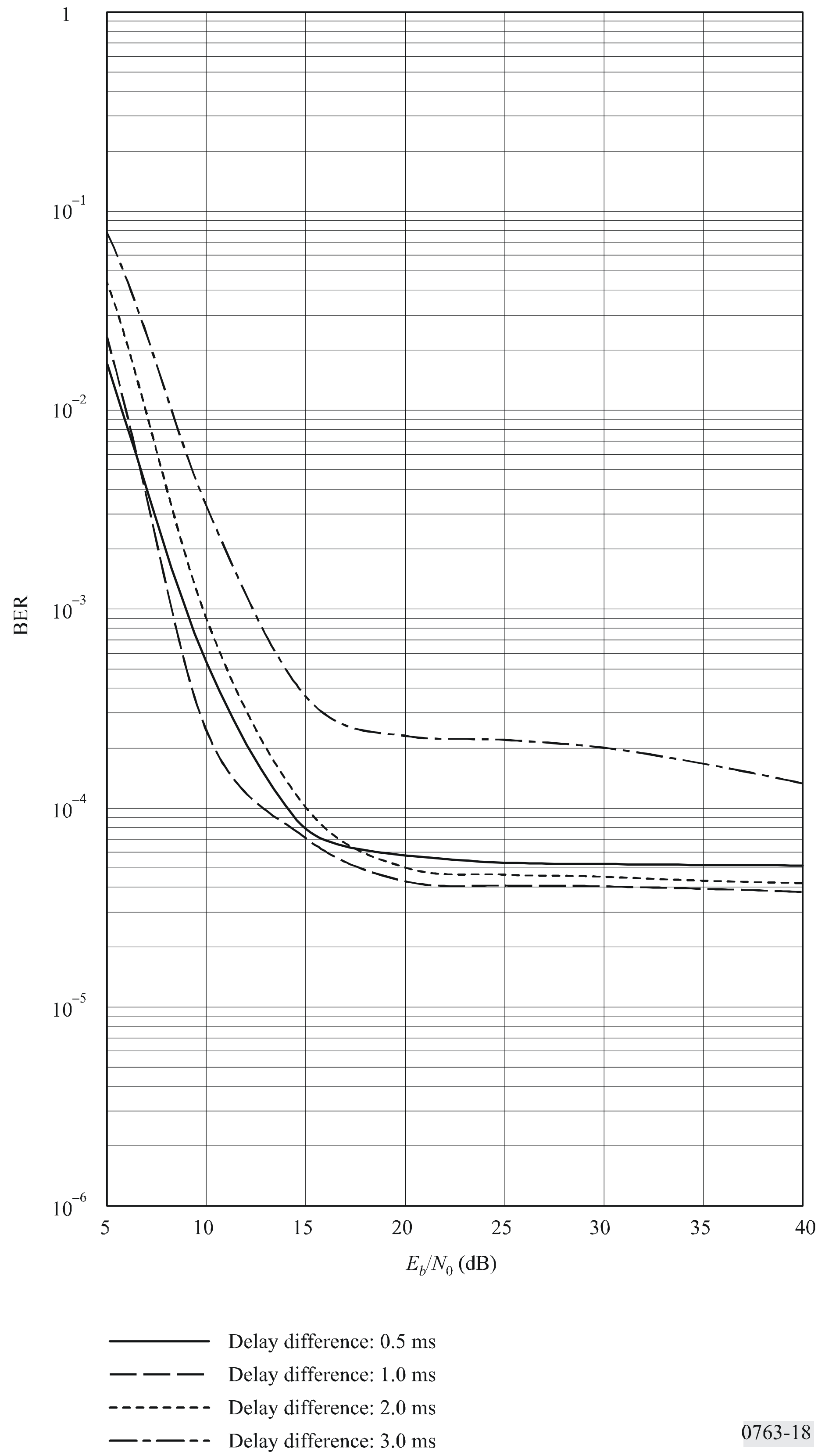
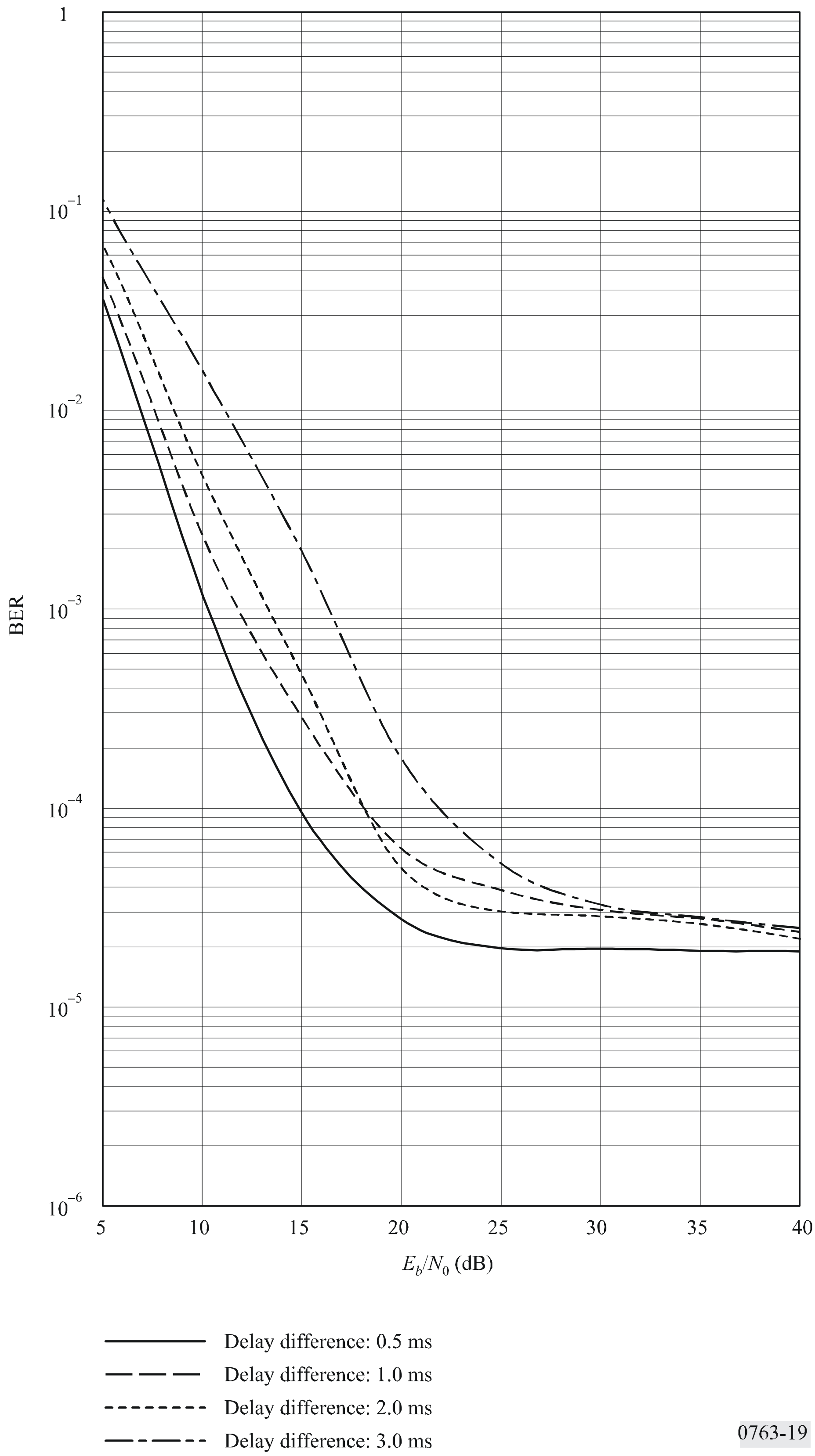


FIGURE 19

BPSK BER versus noise spectral density for a fading channel



Annex 6

High data rate waveforms 3 200/4 800/6 400/8 000/9 600/12 800 bit/s using a serial transmission modem over HF circuits

1 Introduction

This Annex provides a detailed description of modem waveforms to ensure operation within HF radio networks. This family of waveforms is also known as STANAG 4539. A family of self-identifying waveforms is described for coded operation from 3 200 bit/s to 9 600 bit/s (with optional uncoded operation at 12 800 bit/s). The self-identifying feature[1] of this family of waveforms enables rapid adaptation of the modulation to respond to changing channel conditions. The key features of this waveform are:

– Ability to track an HF channel with 3-5 ms of multipath fading.
– Ability to correct for errors caused by fading, multipath and noise.
– The equipment passband bandwidth requirement is 300 to 3 050 Hz.
– Automatic data rate and interleaver detection.
– Able to tolerate a shift of ±75 Hz between the transmission and reception HF carriers.

1.1 Overview

This section presents a modem waveform and coding for data rates of 3 200, 4 800, 6 400, 8 000, 9 600 and uncoded optional operation at 12 800 bit/s.

A block interleaver is used to obtain six interleaving lengths ranging from 0.12 s to 8.64 s. A single coding option, a constraint length 7, rate 1/2 convolutional code, punctured to rate 3/4, is used for all data rates. The full-tail-biting approach is used to produce block codes from this convolutional code that are the same length as the interleaver.

Both the data rate and interleaver settings are explicitly transmitted as a part of the waveform, both as part of the initial preamble and then periodically as both a reinserted preamble and in the periodic known symbol blocks. This self-identifying feature is important in developing efficient (ARQ) protocols for HF channels. The receive modem is able to deduce the data rate and interleaver setting either from the preamble or from the subsequent data portion of the waveform.

1.2 Modulation

The symbol rate for all symbols is 2 400 symbols/s, which should be accurate to a minimum of ±0.024 symbols/s (10 ppm) when the transmit data clock is generated by the modem and not provided by the data terminal equipment (DTE). PSK and QAM modulation techniques are used. The sub-carrier (or pair of quadrature sub-carriers in the case of QAM) is centred at 1 800 Hz accurate to 0.018 Hz (10 ppm). The phase of the quadrature sub-carrier relative to the in-phase carrier is 90º. The power spectral density of the modulator output signal is constrained to be at least 20 dB below the signal level measured at 1 800 Hz, when tested outside of the band from 200 Hz to 3 400 Hz. The filter employed should introduce a ripple of no more than ±2 dB in the range from 800 Hz to 2 800 Hz. The filter used is a square root Nyquist filter with $\alpha = 0.35$.

[1] Symbols sent in the preamble and channel probe phases specify data rate and interleaver depth.

1.2.1 Known symbols

For all known symbols, the modulation used is PSK, with the symbol mapping shown in Table 3 and Fig. 20. No scrambling is applied to the known symbols.

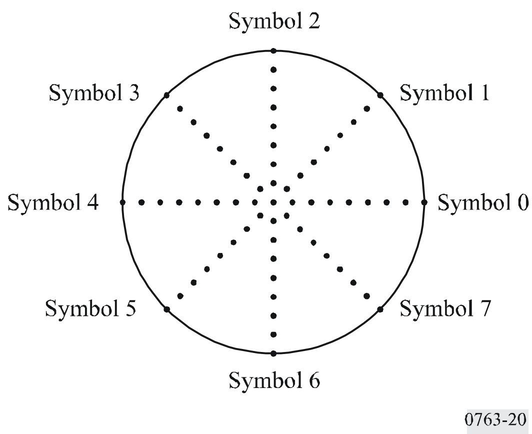
TABLE 3

8-PSK symbol mapping

Symbol number	Phase	In-phase	Quadrature
0	0	1.000000	0.000000
1	$\pi/4$	0.707107	0.707107
2	$\pi/2$	0.000000	1.000000
3	$3\pi/4$	–0.707107	0.707107
4	π	–1.000000	0.000000
5	$5\pi/4$	–0.707107	–0.707107
6	$3\pi/2$	0.0000000	–1.000000
7	$7\pi/4$	0.707107	–0.707107

FIGURE 20

8-PSK signal constellation and symbol mapping



1.2.2 Data symbols

For data symbols, the modulation used will depend upon the data rate. Table 4 describes the modulation that is used with each data rate.

TABLE 4

Modulation used to obtain each data rate

Data rate (bit/s)	Modulation
3 200	QPSK
4 800	8-PSK
6 400	16-QAM
8 000	32-QAM
9 600	64-QAM
12 800	64-QAM

Both the 16-QAM and 32-QAM constellations employ multiple PSK rings to maintain good peak-to-average ratios, and the 64-QAM constellation is a variation of the standard square QAM constellation, which has been modified to improve the peak-to-average ratio.

1.2.2.1 PSK data symbols

For the PSK constellations, a distinction is made between the data bits and the symbol number for the purposes of scrambling the QPSK modulation to appear as 8-PSK, on-air. Scrambling is applied as a modulo 8 addition of a scrambling sequence to the 8-PSK symbol number. Transcoding is an operation which links a symbol to be transmitted to a group of data bits.

1.2.2.1.1 QPSK symbol mapping

For the 3 200 bit/s user data rate, transcoding is achieved by linking one of the symbols specified in Table 3 to a set of two consecutive data bits (dibit) as shown in Table 5. In this Table, the leftmost bit of the dibit is the older bit; i.e. fetched from the interleaver before the rightmost bit.

TABLE 5

Transcoding for 3 200 bit/s

Dibit	Symbol
00	0
01	2
11	4
10	6

1.2.2.1.2 8-PSK symbol mapping

For the 4 800 bit/s user data rate, transcoding is achieved by linking one symbol to a set of three consecutive data bits (tribit) as shown in Table 6. In this Table, the leftmost bit of the tribit is the oldest bit; i.e. fetched from the interleaver before the other two, and the rightmost bit is the most recent bit.

TABLE 6

Transcoding for 4 800 bit/s

Tribit	Symbol
000	1
001	0
010	2
011	3
100	6
101	7
110	5
111	4

1.2.2.1.3 QAM data symbols

For the QAM constellations, no distinction is made between the number formed directly from the data bits and the symbol number. Each set of 4 bits (16-QAM), 5 bits (32-QAM) or 6 bits (64-QAM) is mapped directly to a QAM symbol. For example, the four bit grouping 0111 would map to symbol 7 in the 16-QAM constellation while the 6 bit grouping 100011 would map to symbol 35 in the 64-QAM constellation. Again, in each case the leftmost bit is the oldest bit, i.e. fetched from the interleaver before the other bits, and the rightmost bit is the most recent bit.

The mapping of bits to symbols for the QAM constellations has been selected to minimize the number of bit errors incurred when errors involve adjacent signalling points in the constellation.

1.2.2.1.4 The 16-QAM constellation

The constellation points, which are for 16-QAM, are shown in Fig. 21 and described in terms of their in-phase and quadrature components in Table 7. As can be seen in the Figure, the 16-QAM constellation comprises two PSK rings: 4-PSK inner and 12-PSK outer symbols.

FIGURE 21

16-QAM signalling constellation

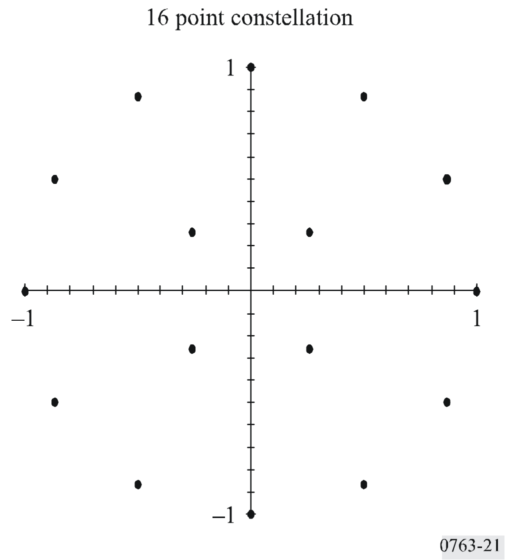


TABLE 7

In-phase and quadrature components of each 16-QAM symbol

Symbol number	In-phase	Quadrature
0	0.866025	0.500000
1	0.500000	0.866025
2	1.000000	0.000000
3	0.258819	0.258819
4	–0.500000	0.866025
5	0.000000	1.000000
6	–0.866025	0.500000
7	–0.258819	0.258819
8	0.500000	–0.866025
9	0.000000	–1.000000
10	0.866025	–0.500000
11	0.258819	–0.258819
12	–0.866025	–0.500000
13	–0.500000	–0.866025
14	–1.000000	0.000000
15	–0.258819	–0.258819

1.2.2.1.5 The 32-QAM constellation

The constellation points, which are used for 32-QAM, are shown in Fig. 22 and specified in terms of their in-phase and quadrature components in Table 8. This constellation contains an outer ring of 16 symbols and an inner square of 16 symbols.

FIGURE 22

32-QAM signalling constellation

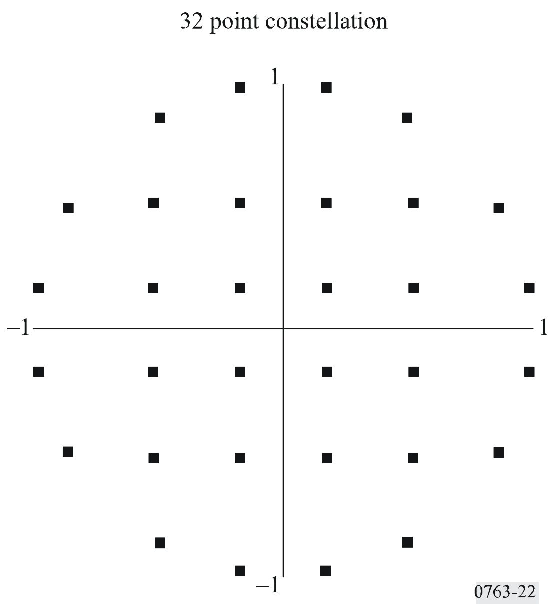


TABLE 8

In-phase and quadrature components of each 32-QAM symbol

Symbol number	In-phase	Quadrature	Symbol number	In-phase	Quadrature
0	0.866380	0.499386	16	0.866380	–0.499386
1	0.984849	0.173415	17	0.984849	–0.173415
2	0.499386	0.866380	18	0.499386	–0.866380
3	0.173415	0.984849	19	0.173415	–0.984849
4	0.520246	0.520246	20	0.520246	–0.520246
5	0.520246	0.173415	21	0.520246	–0.173415
6	0.173415	0.520246	22	0.173415	–0.520246
7	0.173415	0.173415	23	0.173415	–0.173415
8	–0.866380	0.499386	24	–0.866380	–0.499386
9	–0.984849	0.173415	25	–0.984849	–0.173415
10	–0.499386	0.866380	26	–0.499386	–0.866380
11	–0.173415	0.984849	27	–0.173415	–0.984849
12	–0.520246	0.520246	28	–0.520246	–0.520246
13	–0.520246	0.173415	29	–0.520246	–0.173415
14	–0.173415	0.520246	30	–0.173415	–0.520246
15	–0.173415	0.173415	31	–0.173415	–0.173415

1.2.2.1.6 The 64-QAM constellation

The constellation points which are used for the 64-QAM modulation are shown in Fig. 23 and described in terms of their in-phase and quadrature components in Table 9. This constellation is a variation on the standard 8 × 8 square constellation, which achieves a better peak-to-average ratio without sacrificing the very good pseudo-Gray code properties of the square constellation.

FIGURE 23

64-QAM signalling constellation

64 point constellation

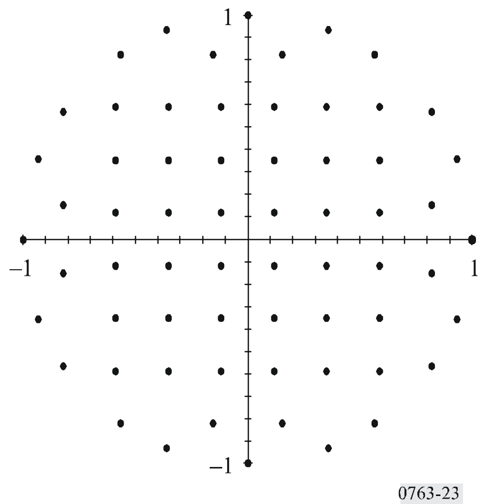


TABLE 9

In-phase and quadrature components of each 64-QAM symbol

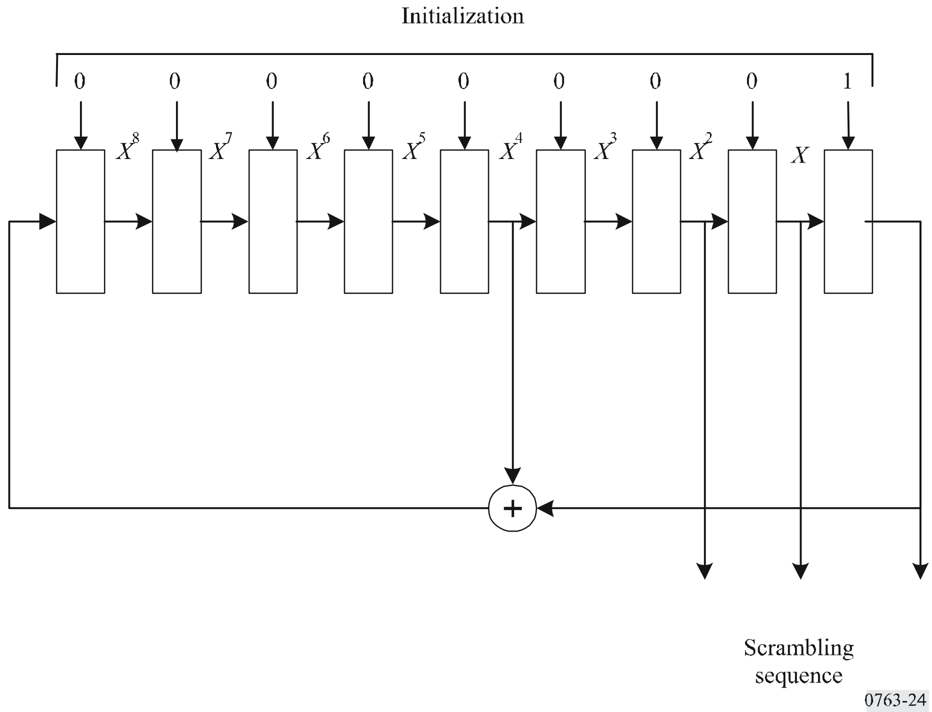
Symbol number	In-phase	Quadrature	Symbol number	In-phase	Quadrature
0	1.000000	0.000000	32	0.000000	1.000000
1	0.822878	0.568218	33	–0.822878	0.568218
2	0.821137	0.152996	34	–0.821137	0.152996
3	0.932897	0.360142	35	–0.932897	0.360142
4	0.000000	–1.000000	36	–1.000000	0.000000
5	0.822878	–0.568218	37	–0.822878	–0.568218
6	0.821137	–0.152996	38	–0.821137	–0.152996
7	0.932897	–0.360142	39	–0.932897	–0.360142
8	0.568218	0.822878	40	–0.568218	0.822878
9	0.588429	0.588429	41	–0.588429	0.588429
10	0.588429	0.117686	42	–0.588429	0.117686
11	0.588429	0.353057	43	–0.588429	0.353057
12	0.568218	–0.822878	44	–0.568218	–0.822878
13	0.588429	–0.588429	45	–0.588429	–0.588429
14	0.588429	–0.117686	46	–0.588429	–0.117686
15	0.588429	–0.353057	47	–0.588429	–0.353057
16	0.152996	0.821137	48	–0.152996	0.821137
17	0.117686	0.588429	49	–0.117686	0.588429
18	0.117686	0.117686	50	–0.117686	0.117686
19	0.117686	0.353057	51	–0.117686	0.353057
20	0.152996	–0.821137	52	–0.152996	–0.821137
21	0.117686	–0.588429	53	–0.117686	–0.588429
22	0.117686	–0.117686	54	–0.117686	–0.117686
23	0.117686	–0.353057	55	–0.117686	–0.353057
24	0.360142	0.932897	56	–0.360142	0.932897
25	0.353057	0.588429	57	–0.353057	0.588429
26	0.353057	0.117686	58	–0.353057	0.117686
27	0.353057	0.353057	59	–0.353057	0.353057
28	0.360142	–0.932897	60	–0.360142	–0.932897
29	0.353057	–0.588429	61	–0.353057	–0.588429
30	0.353057	–0.117686	62	–0.353057	–0.117686
31	0.353057	–0.353057	63	–0.353057	–0.353057

1.2.3 Data scrambling

Data symbols for the 8-PSK symbol constellation (3 200 bit/s, 4 800 bit/s) are scrambled by modulo 8 addition with a scrambling sequence. The data symbols for the 16-QAM, 32-QAM, and 64-QAM constellations are scrambled by using an exclusive or (XOR) operation. Sequentially, the data bits forming each symbol (4 for 16-QAM, 5 for 32-QAM, and 6 for 64-QAM) are XOR'd with an equal number of bits from the scrambling sequence. In all cases, the scrambling sequence generator polynomial is $X^9 + X^4 + 1$ and the generator is initialized to 1 at the start of each data frame. A block diagram of the scrambling sequence generator is shown in Fig. 24.

FIGURE 24

Scrambling sequence generator illustrating scrambling generator for 8-PSK symbols



For 8-PSK symbols (3 200 bit/s and 4 800 bit/s), the scrambling is carried out taking the modulo 8 sum of the numerical value of the binary triplet consisting of the last (rightmost) three bits in the shift register, and the symbol number (transcoded value). For example, if the last three bits in the scrambling sequence shift register were 010 which has a numerical value equal 2, and the symbol number before scrambling was 6, symbol 0 would be transmitted since: (6 + 2) modulo 8 = 0. For 16-QAM symbols, scrambling is carried out by XORing the 4 bit number consisting of the last (rightmost) four bits in the shift register with the symbol number. For example, if the last 4 bits in the scrambling sequence shift register were 0101 and the 16-QAM symbol number before scrambling was 3 (i.e. 0011), symbol 6 (0110) would be transmitted. For 32-QAM symbols, scrambling is carried out by XORing the 5 bit number formed by the last (rightmost) five bits in the shift register with the symbol number. For 64-QAM symbols, scrambling is carried out by XORing the 6 bit number formed by the last (rightmost) six bits in the shift register with the symbol number.

After each data symbol is scrambled, the generator is iterated (shifted) the required number of times to produce all new bits for use in scrambling the next symbol (i.e. three iterations for 8-PSK, four iterations for 16-QAM, five iterations for 32-QAM and six iterations for 64-QAM). Since the generator is iterated after the bits are used, the first data symbol of every data frame should be scrambled by the appropriate number of bits from the initialization value of 00000001.

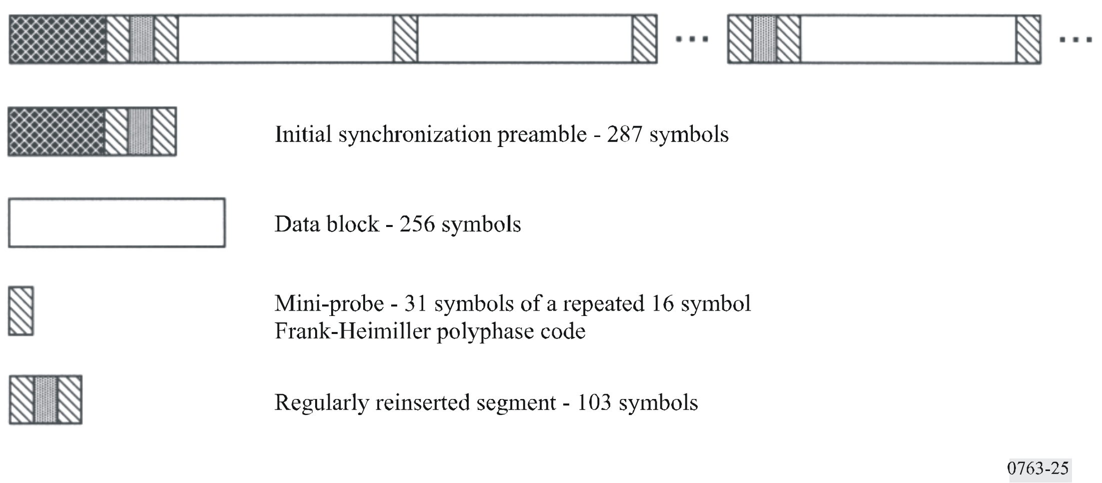
The length of the scrambling sequence is 511 bits. For a 256 symbol data block with 6 bits per symbol, this means that the scrambling sequence will be repeated just slightly more than three times, although in terms of symbols, there will be no repetition.

1.3 Frame structure

The frame structure that is used for the waveforms of this Annex is shown in Fig. 25. An initial 287 symbol preamble is followed by 72 frames of alternating data and known symbols. Each data frame has a data block consisting of 256 data symbols, followed by a mini-probe of 31 symbols of known data. After 72 data frames, a 72 symbol subset of the initial preamble is reinserted to facilitate late acquisition, Doppler shift removal, and synchronization adjustment. The total length of known data in this segment is actually 103 symbols: the 72 reinserted preamble symbols plus the preceding 31 symbol mini-probe segment which follows the last 256 symbol data block.

FIGURE 25

Frame structure for all waveforms



1.3.1 Synchronization and reinserted preambles

The synchronization preamble is used for rapid initial synchronization. The reinserted preamble is used to facilitate acquisition of an ongoing transmission (acquisition on data).

1.3.1.1 Synchronization preamble

The synchronization preamble consists of two parts. The first part consists of at least N blocks of 184 8-PSK symbols to be used exclusively for radio and modem AGC. The value of N is configurable to range from values of 0 to 7 (for $N = 0$ this first section is not sent at all). These 184 symbols are formed by taking the complex conjugate of the first 184 symbols of the sequence specified below for the second section.

The second section consists of 287 symbols. The first 184 symbols are intended exclusively for synchronization and Doppler offset removal purposes while the final 103 symbols, which are common with the reinserted preamble, also carry information regarding the data rate and interleaver settings. Expressed as a sequence of 8-PSK symbols, using the symbol numbers given in Table 3 the second section of the synchronization preamble is as follows:

1, 5, 1, 3, 6, 1, 3, 1, 1, 6, 3, 7, 7, 3, 5, 4, 3, 6, 6, 4, 5, 4, 0, 2, 2, 2, 6, 0, 7, 5, 7, 4, 0, 7, 5, 7, 1, 6, 1, 0, 5, 2, 2, 6, 2, 3, 6, 0, 0, 5, 1, 4, 2, 2, 2, 3, 4, 0, 6, 2, 7, 4, 3, 3, 7, 2, 0, 2, 6, 4, 4, 1, 7, 6, 2, 0, 6, 2, 3, 6, 7, 4, 3, 6, 1, 3, 7, 4, 6, 5, 7, 2, 0, 1, 1, 1, 4, 4, 0, 0, 5, 7, 7, 4, 7, 3, 5, 4, 1, 6, 5, 6, 6, 4, 6, 3, 4, 3, 0, 7, 1, 3, 4, 7, 0, 1, 4, 3, 3, 3, 5, 1, 1, 1, 4, 6, 1, 0, 6, 0, 1, 3, 1, 4, 1, 7, 7, 6, 3, 0, 0, 7, 2, 7, 2, 0, 2, 6, 1, 1, 1, 2, 7, 7, 5, 3, 3, 6, 0, 5, 3, 3, 1, 0, 7, 1, 1, 0, 3, 0, 4, 0, 7, 3, 0, 0, 0, 0, 0, 2, 4, 6, 0, 4, 0, 4, 0, 6, 4, 2, 0, 0, 0, 0, 0, 2, 4, 6, 0, 4, 0, 4, 0, 6, 4, 2, (D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0 + 0, 4, 0, 4, 0, 0, 4, 4, 0, 0, 0, 0, 0) Modulo 8 (D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1 + 0, 4, 0, 4, 0, 0, 4, 4, 0, 0, 0, 0, 0) Modulo 8 (D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2 + 0, 4, 0, 4, 0, 0, 4, 4, 0, 0, 0, 0, 0) Modulo 8 6, 4, 4, 4, 4, 4, 6, 0, 2, 4, 0, 4, 0, 4, 2, 0, 6, 4, 4, 4, 4, 4, 6, 0, 2, 4, 0, 4, 0, 4, 2, 0.

where the data symbols D_0, D_1 and D_2 take one of 30 sets of values chosen from Table 10 to indicate the data rate and interleaver settings. The modulo operations are meant to signify that each of the D values are used to shift the phase of a length 13 bit Barker code (0101001100000) by performing modulo 8 addition of the D value with each of the Barker code 13 phase values (0 or 4). This operation can encode 6 bits of information using QPSK modulation of the 13 bit (chip) Barker codes. Since the three Barker code sequences only occupy 39 symbols, the 31 symbol mini-probes are lengthened to 32 symbols each to provide the additional two symbols required to pad the three 13 symbol Barker codes up to a total of 41 symbols.

TABLE 10

D_0, D_1, D_2 8-PSK symbol values as a function of data rate and interleaver length

Data rate (bit/s)	Interleaver length in frames (256 symbol data blocks)					
	1	3	9	18	36	72
3 200	0,0,4	0,2,6	0,2,4	2,0,6	2,0,4	2,2,6
4 800	0,6,2	0,4,0	0,4,2	2,6,0	2,6,2	2,4,0
6 400	0,6,4	0,4,6	0,4,4	2,6,6	2,6,4	2,4,6
8 000	6,0,2	6,2,0	6,2,2	4,0,0	4,0,2	4,2,0
9 600	6,0,4	6,2,6	6,2,4	4,0,6	4,0,4	4,2,6
12 800	6,6,2[(1)]	N/A	N/A	N/A	N/A	N/A

N/A: Not applicable

[(1)] For 12 800 bit/s 1 frame interleaver interpreted as no interleaving.

The mapping chosen to create Table 10 uses 3 bits each to specify the data rate and interleaver length. The 3 data rate bits are the three most significant bits (MSB) of 3 dibit symbols and the interleaver length bits are the least significant bits (LSB). The phase of the Barker code is determined from the three resulting dibit words using Table 5, the dibit transcoding table. The 3 bit data rate and interleaver length mappings are shown in Table 11. Note that the transcoding has the effect of placing the 3 interleaver length bits in quadrature with the 3 data rate bits.

TABLE 11

Bit patterns for specifying data rate and interleaver length

Data rate	3 Bit mapping
3 200	001
4 800	010
6 400	011
8 000	100
9 600	101
12 800	110

Interleaver length	3 Bit mapping	Name
1 Frame	001	Ultra short (US)
3 Frames	010	Very short (VS)
9 Frames	011	Short (S)
18 Frames	100	Medium (M)
36 Frames	101	Long (L)
72 Frames	110	Very long (VL)

Since the Barker code is unbalanced in terms of the number of 0s and 1s, these 3-bit patterns have been chosen to avoid the 000 or 111 patterns in order to minimize the unbalance in the combined three symbols. More specifically, one of the three repeats of the Barker code that appears on each of the quadrature components is always phase shifted by 180° with respect to the other two. This results in a net imbalance in each quadrature component of the 39 symbols that is always 17 to 22, rather than ever being 12 to 27.

1.3.1.2 Reinserted preamble

The reinserted preamble is identical to the final 72 symbols of the synchronization preamble. In fact, the final 103 symbols are common between the synchronization preamble and the contiguous block consisting of the reinserted preamble and the mini-probe which immediately precedes it. The 103 symbols of known data (including the 31 mini-probe symbols of the preceding data frame) are thus:

0, 0, 0, 0, 0, 2, 4, 6, 0, 4, 0, 4, 0, 6, 4, 2, 0, 0, 0, 0, 0, 2, 4, 6, 0, 4, 0, 4, 0, 6, 4,
2,
(D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0, D_0 + 0, 4, 0, 4, 0, 0, 4, 4, 0, 0, 0, 0, 0) Modulo 8
(D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1, D_1 + 0, 4, 0, 4, 0, 0, 4, 4, 0, 0, 0, 0, 0) Modulo 8
(D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2, D_2 + 0, 4, 0, 4, 0, 0, 4, 4, 0, 0, 0, 0, 0) Modulo 8
6,
4, 4, 4, 4, 4, 6, 0, 2, 4, 0, 4, 0, 4, 2, 0, 6, 4, 4, 4, 4, 4, 6, 0, 2, 4, 0, 4, 0, 4, 2, 0.

where the data symbols D_0, D_1, and D_2 again take one of 30 sets of values chosen from Table 10 to indicate the data rate and interleaver settings as described in § 1.3.1.1. Note that the first 31 of these symbols are the immediately preceding mini-probe, which follows the last of the 72 data blocks.

1.3.2 Mini-probes

Mini-probes 31 symbols in length are inserted following every 256 symbol data block and at the end of each preamble (where they are considered to be part of the preamble). Using the 8-PSK symbol mapping, each mini-probe is based on the repeated Frank-Heimiller sequence. The sequence that is used, specified in terms of the 8-PSK symbol numbers, is given by:

0, 0, 0, 0, 0, 2, 4, 6, 0, 4, 0, 4, 0, 6, 4, 2, 0, 0, 0, 0, 0, 2, 4, 6, 0, 4, 0, 4, 0, 6, 4.

This mini-probe will be designated "+". The phase inverted version of this is:

4, 4, 4, 4, 4, 6, 0, 2, 4, 0, 4, 0, 4, 2, 0, 6, 4, 4, 4, 4, 4, 6, 0, 2, 4, 0, 4, 0, 4, 2, 0.

and mini-probes using this sequence will be designated "–", as the phase of each symbol has been rotated 180º from the "+".

There are a total of 73 mini-probes for each set of 72 data blocks. For convenience, each mini-probe is sequentially numbered, with mini-probe 0 being defined as the last 31 symbols of the preceding (reinserted) preamble, mini-probe number 1 following the first data block after a (reinserted) preamble. Mini-probe 72 follows the 72nd data block, and is also the first 31 symbols of the next 103 symbol reinserted preamble. Mini-probes 0 and 72 have been defined as part of the reinsertion preamble to have the signs – and + respectively. The data rate and interleaver length information encoded into the synchronization and reinserted preambles are also be encoded into mini-probes 1 to 72. These 72 mini-probes are grouped into four sets of 18 consecutive mini-probes (1 to 18, 19 to 36, 37 to 54, and 55 to 72). Note that the 256 symbol data block that immediately follows the 18th mini-probe, in each of the first three sets, is also the 1st data block of an interleaver block with frame lengths of 1, 3, 9, and 18. The length 36 interleaver block begins after the second set, and a reinserted preamble begins after the fourth set. This structure permits data to begin to be demodulated as soon as the interleaver boundary becomes known.

Each 18 mini-probe sequence consists of seven – signs, a + sign, followed by six sign values that are dependent on the data rate and interleaver length, three sign values that specify which of the four sets of 18 mini-probes it is, and then finally a + sign. For the fourth set, this final + sign (mini-probe 72) is also the initial mini-probe of the next reinserted preamble (which uses the + phase).

Pictorially, this length 18 sequence is: – – – – – – – + S_0 S_1 S_2 S_3 S_4 S_5 S_6 S_7 S_8 +, where the first six S_i sign values are defined in Table 12. Note that these 6 bit patterns (+ is a 0) correspond to the concatenation of the 3 bit mappings from Table 11 for the data rate (S_0 S_1 S_2) and the interleaver length (S_3 S_4 S_5). The final three S_i sign values which specify the mini-probe set (count) are defined in Table 13.

TABLE 12

S_0, S_1, S_2, S_3, S_4, S_5 (sign) values as a function of data rate and interleaver setting

Data rate (bit/s)	**Interleaver length in frames (256 symbol data blocks)**					
	1	**3**	**9**	**18**	**36**	**72**
3 200	+ + – + + –	+ + – + – +	+ + – + – –	+ + – – + +	+ + – – + –	+ + – – – +
4 800	+ – + + + –	+ – + + – +	+ – + + – –	+ – + – + +	+ – + – + –	+ – + – – +
6 400	+ – – + + –	+ – – + – +	+ – – + – –	+ – – – + +	+ – – – + –	+ – – – – +
8 000	– + + + + –	– + + + – +	– + + + – –	– + + – + +	– + + – + –	– + + – – +
9 600	– + – + + –	– + – + – +	– + – + – –	– + – – + +	– + – – + –	– + – – – +
12 800	– – + + + –	N/A	N/A	N/A	N/A	N/A

TABLE 13

S_6, S_7, S_8 (sign) values as a function of mini-probe set

Mini-probe set			
1 to 18	**19 to 36**	**37 to 54**	**55 to 72**
+ + –	+ – +	+ – –	– + +

The first eight mini-probes in each set (– – – – – – – +) uniquely locate the starting point for the following nine S_i values. This is possible since the S_i sequences used contain at most runs of four + or – phases. This makes it impossible for a sequence of seven mini-probes with the same phase followed by one with a phase reversal to occur anywhere else except at the beginning of one of the 18 mini-probe sequences. Once this fixed eight mini-probe pattern is located, the 0º or 180º phase ambiguity is also resolved so that the following nine mini-probes can be properly matched to the data rate, interleaver length, and mini-probe set count. The entire mini-probe sequence is as follows:

[rp] – – – – – – – + S_0 S_1 S_2 S_3 S_4 S_5 S_6 S_7 S_8 + – – – – – – – + S_0 S_1 S_2 S_3 S_4 S_5 S_6 S_7 S_8 + – – – – – – – + S_0 S_1 S_2 S_3 S_4 S_5 S_6 S_7 S_8 + – – – – – – – + S_0 S_1 S_2 S_3 S_4 S_5 S_6 S_7 S_8 [rp]

where the [rp] represents the 103 reinserted preamble symbols (includes mini-probes 72 and 0).

1.4 Coding and interleaving

The interleaver is a block interleaver. Each block of input data is also encoded using a block encoding technique with a code block size equal to the size of the block interleaver. Thus, the input data bits will be sent as successive blocks of bits that span the duration of the interleaver length selected. Table 14 shows the number of input data bits per block as function of both data rate and interleaver length. Note that an "input data block" should not be confused with the 256 symbol data block that is part of a data frame in the waveform format. The bits from an input data block will be mapped through the coding and interleaving to the number of data frames, and thus 256 symbol data blocks, that define the interleaver length.

TABLE 14

Input data block size in bits as a function of data rate and interleaver length

Data rate (bit/s)	Interleaver length in frames					
	1	3	9	18	36	72
	Number of input data bits per block					
3 200	384	1 152	3 456	6 912	13 824	27 648
4 800	576	1 728	5 184	10 368	20 736	41 472
6 400	768	2 304	6 912	13 824	27 648	55 296
8 000	960	2 880	8 640	17 280	34 560	69 120
9 600	1 152	3 456	10 368	20 736	41 472	82 944

1.4.1 Block boundary alignment

Each code block is interleaved within a single interleaver block of the same size. The boundaries of these blocks are aligned such that the beginning of the first data frame following each reinserted preamble should coincide with an interleaver boundary. Thus for an interleaver length of three frames, the first three data frames following a reinserted preamble will contain all of the encoded bits for a single input data block. The first data symbol from the first data frame in each interleaver set will have as its MSB the first bit fetched from the interleaver.

1.4.2 Block encoding

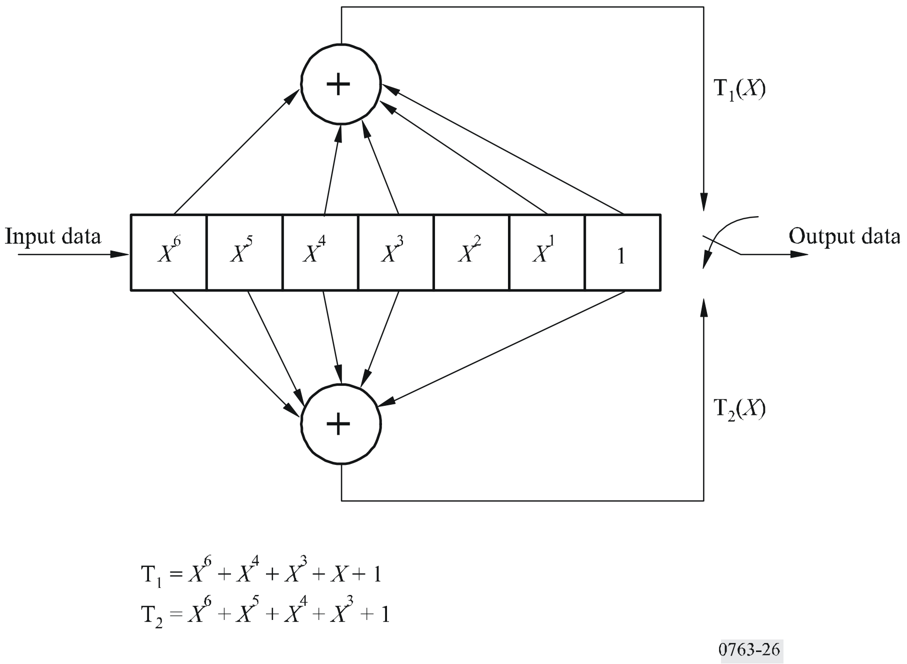
The full-tail-biting and puncturing techniques is used with a rate 1/2 convolutional code to produce a rate 3/4 block code that is the same length as the interleaver.

1.4.3 Rate 1/2 convolutional code

A constraint length 7, rate 1/2 convolutional code is used prior to puncturing. Figure 26 is a pictorial representation of the encoder. The two generator polynomials used are:

FIGURE 26

Constraint length 7, rate 1/2 convolutional encoder



The two summing nodes in the Figure represent modulo 2 addition. For each bit input to the encoder, two bits are taken from the encoder, with the upper output bit, $T_1(X)$, taken first.

1.4.3.1 Full-tail-biting encoding

To begin encoding each block of input data, the encoder is preloaded by shifting in the first six input data bits without taking any output bits. These six input bits is temporarily saved so that they can be used to "flush" the encoder. The first two coded output bits are taken after the seventh bit has been shifted in, and are defined to be the first two bits of the resulting block code. After the last input data bit has been encoded, the first six "saved" data bits are encoded. Note that the encoder shift register should not be changed before encoding these saved bits; i.e. it should be filled with the last seven input data bits. The six "saved" data bits are encoded by shifting them into the encoder one at a time, beginning with the earliest of the six. The encoding thus continues by taking the two resulting coded output bits as each of the saved six bits is shifted in. These encoded bits are the final bits of the resulting (unpunctured) block code. Prior to puncturing, the resulting block code will have exactly twice as many bits as the input information bits. Puncturing of the rate 1/2 code to the required rate 3/4 is done prior to sending bits to the interleaver.

1.4.3.2 Puncturing to rate 3/4

In order to obtain a rate 1/2 code from the rate 3/4 code used, the output of the encoder must be punctured by not transmitting 1 bit out of every 3. Puncturing is performed by using a puncturing mask of 1 1 1 0 0 1, applied to the bits output from the encoder. In this notation a 1 indicates that the bit is retained and a 0 indicates that the bit is not transmitted. For an encoder generated sequence of:

$$T_1(k),\ T_2(k),\ T_1(k+1),\ T_2(k+1),\ T_1(k+2),\ T_2(k+2)\ ...$$

the transmitted sequence would be:

$$T_1(k), T_2(k), T_1(k+1), T_1(k+2) \ldots$$

Defining $T_1(0)$, $T_2(0)$ to be the first two bits of the block code generated as defined in § 1.4.2, then the value of k in the above sequences is an integral multiple of 3. The block code is punctured in this manner before being input to the interleaver.

1.4.4 Block interleaver structure

The block interleaver used is designed to separate neighbouring bits in the punctured block code as far as possible over the span of the interleaver with the largest separations resulting for the bits that were originally closest to each other. Because of the 30 different combinations of data rates and interleaver lengths, a flexible interleaver structure is needed.

1.4.4.1 Interleaver size in bits

The interleaver consists of a single dimension array, numbered from 0 to its size in bits –1. The array size depends on both the data rate and interleaver length selected as shown in Table 15.

TABLE 15

Interleaver size in bits as a function of data rate and interleaver length

Data rate (bit/s)	Interleaver length in frames					
	1	3	9	18	36	72
	Interleaver size in bits					
3 200	512	1 536	4 608	9 216	18 432	36 864
4 800	768	2 304	6 912	13 824	27 648	55 296
6 400	1 024	3 072	9 216	18 432	36 864	73 728
8 000	1 280	3 840	11 520	23 040	46 080	92 160
9 600	1 536	4 608	13 824	27 648	55 296	110 592

1.4.4.2 Interleaver load

The punctured block code bits are loaded into the interleaver array beginning with location 0. The location for loading each successive bit is obtained from the previous location by incrementing by the "interleaver increment value" specified in Table 16, modulo the "interleaver size in bits".

Defining the first punctured block code bit to be B(0), then the load location for B(n) is given by:

Load location = (n * Interleaver increment value) Modulo (Interleaver size in bits)

Thus for 3 200 bit/s, with a one frame interleaver (512 bit size with an increment of 97), the first 8 interleaver load locations are: 0, 97, 194, 291, 388, 485, 70, and 167.

TABLE 16

Interleaver increment value as a function of data rate and interleaver length

Data rate (bit/s)	Interleaver length in frames					
	1	3	9	18	36	72
	Interleaver increment value					
3 200	97	229	805	1 393	3 281	6 985
4 800	145	361	1 045	2 089	5 137	10 273
6 400	189	481	1 393	3 281	6 985	11 141
8 000	201	601	1 741	3 481	8 561	14 441
9 600	229	805	2 089	5 137	10 273	17 329

These increment values have been chosen to ensure that the combined cycles of puncturing and assignment of bit positions in each symbol for the specific constellation being used is the same as if there had been no interleaving. This is important, because each symbol of a constellation contains "strong" and "weak" bit positions, except for the lowest data rate. Bit position refers to the location of the bit, ranging from MSB to LSB, in the symbol mapping. A strong bit position is one that has a large average distance between all the constellation points where the bit is a 0 and the closest point where it is a 1. Typically, the MSB is a strong bit and the LSB a weak bit. An interleaving strategy that did not evenly distribute these bits in the way they occur without interleaving could degrade performance.

1.4.4.3 Interleaver fetch

The fetching sequence for all data rates and interleaver lengths start with location 0 of the interleaver array and increment the fetch location by 1. This is a simple linear fetch from beginning to end of the interleaver array.

1.5 Operational features and message protocols

The format of this high data rate waveform has been designed to permit it to work well with most of the protocols used and planned for use with HF. The reinserted preamble facilitates acquisition (or reacquisition) of an ongoing broadcast transmission. The short length of the synchronization preamble, wide range of interleaving lengths, and the use of full-tail-biting coding are intended to provide efficient operation with ARQ protocols. To further enhance the operation with these protocols, the following operational features are included in the HF modem.

1.5.1 Onset of transmission

The modem begins a transmission no later than 100 ms after it has received an entire input data block (enough bits to fill a coded and interleaved block), or upon receipt of the last input data bit, whichever occurs first. The latter would only occur when the message is shorter than one interleaver block. A transmission is defined as beginning with the keying of the radio, followed by the output of the preamble waveform after the configured pre-key delay, if any.

The delay between when the modem receives the first input data bit and the onset of transmission will be highly dependent on the means for delivery of the input data bits to the modem. A synchronous serial interface at the user data rate will have the greatest delay. For this reason it is advisable that a high-speed asynchronous interface (serial or Ethernet port) with flow-control be used if this delay is of concern for the particular application.

1.5.2 End of message

The use of an end-of-message (EOM) in the transmit waveform is a configurable option. When the use of an EOM has been selected, a 32-bit EOM pattern is appended after the last input data bit of the message. The EOM, expressed in hexadecimal notation is 4B65A5B2, where the left most bit is sent first. If the last bit of the EOM does not fill out an input data block, the remaining bits in the input data block is set to zero before encoding and interleaving the block.

If the use of an EOM has been inhibited, and the last input data bit does not fill out an input data block, the remaining bits in the input data block is set to zero before encoding and interleaving the block. It is anticipated that the use of an EOM will only be inhibited when an ARQ data protocol uses ARQ blocks which completely fill (or nearly so) the selected input data block size (interleaver block). Without this feature, the use of an EOM would require the transmission of an additional interleaver block under these circumstances.

1.5.3 Termination of a transmission

The modem should terminate a transmission only after the transmission of the final data frame, including a mini-probe, associated with the final interleaver block. Note that a data frame consists of a 256 symbol data block followed by a mini-probe. Also any signal processing and/or filter delays in the modem and the HF transmitter must be accounted for (as part of the key line control timing) to ensure that the entire final mini-probe is transmitted before the transmitter power is turned off.

1.5.4 Termination of receive data processing

There are a number of events, which can cause the HF modem to cease processing the received signal to recover data, and return to the acquisition mode. These are necessary because a modem is not able to acquire a new transmission while it is attempting to demodulate and decode data.

1.5.4.1 Detection of EOM

The HF modem should always scan all of the decoded bits for the 32-bit EOM pattern defined in § 1.5.2. Upon detection of the EOM, the modem will return to the acquisition mode. The modem should continue to deliver decoded bits to the user (DTE) until the final bit immediately preceding the EOM has been delivered.

1.5.4.2 Receipt of a specified number of data blocks

The maximum message duration measured in number of input data blocks (interleaver blocks) is a configurable parameter. Setting this parameter to zero will specify that an unlimited number may be received. Once the modem has decoded and delivered to the user (DTE), the number of bits corresponding to the configured maximum message duration, the HF modem should return to the acquisition mode and terminate the delivery of decoded bits to the user (DTE). Operation with a specified number of input data blocks may be used by an ARQ protocol where the size of the ARQ packet is fixed, or occasionally changed to accommodate changing propagation conditions. In this case, it is anticipated that this parameter (maximum message duration) should be sent to the receiving end of the link as part of the ARQ protocol. It would then be sent to the receiving modem through the remote control interface since it is not embedded in the waveform itself as the data rate and interleaver length parameters are.

1.6 Performance capabilities

The performance capabilities for the high data rate mode are presented in this section. These test results demonstrate that the modem operates reliably over HF circuits for the channel impairments tested.

1.6.1 Simulator characteristics

The high data rate mode was tested using a baseband HF simulator patterned after the Watterson Model in accordance with Recommendation ITU-R F.1487. Additive white Gaussian noise (AWGN) was used as the noise source. Both signal and noise power was measured in a 3 kHz bandwidth.

1.6.2 Radio filters

Finite impulse response (FIR) filters that reflect radio passband requirements were used. The filter is an $N = 63$ FIR filter with the following coefficients (read across, then down) and has a sample rate of 16 000 samples/s:

3.4793306E-04	–4.6615634E-05	3.6863006E-05	6.8983925E-04
1.2186785E-03	7.1322870E-04	–6.2685051E-04	–1.1305640E-03
3.8082659E-04	2.2257954E-03	1.0150929E-03	–3.6258003E-03
–6.9094691E-03	–4.2534569E-03	1.1371180E-03	–1.0868903E-04
–1.1312117E-02	–2.2036370E-02	–1.8856425E-02	–4.9115933E-03
–1.3025356E-03	–2.1579735E-02	–4.8379221E-02	–4.8040411E-02
–1.4815010E-02	9.8565688E-03	–2.0275153E-02	–9.0223589E-02
–1.1587973E-01	–2.2672007E-02	1.6315786E-01	3.1537800E-01
3.1537800E-01	1.6315786E-01	–2.2672007E-02	–1.1587973E-01
–9.0223589E-02	–2.0275153E-02	9.8565688E-03	–1.4815010E-02
–4.8040411E-02	–4.8379221E-02	–2.1579735E-02	–1.3025356E-03
–4.9115933E-03	–1.8856425E-02	–2.2036370E-02	–1.1312117E-02
–1.0868903E-04	1.1371180E-03	–4.2534569E-03	–6.9094691E-03
–3.6258003E-03	1.0150929E-03	2.2257954E-03	3.8082659E-04
–1.1305640E-03	–6.2685051E-04	7.1322870E-04	1.2186785E-03
6.8983925E-04	3.6863006E-05	–4.6615634E-05	3.4793306E-04

1.6.3 BER performance

BER performance was measured using radio filters, with the HF channel simulator programmed to simulate the following channels for a BER of 1×10^{-4}:

– The AWGN channel consists of a single, non-fading path. Each condition was measured for 15 min.

– The Rician channel consists of two independent but equal average power paths, with a fixed 2 ms delay between paths. The first path was non-fading. The second was a Rayleigh fading path with a two sigma fading BW of 2 Hz. Each condition was measured for 2 h.

– The Recommendation ITU-R F.1487 “mid latitudes disturbed” condition (Poor channel) consists of two independent but equal average power Rayleigh fading paths, with a fixed 2 ms delay between paths, and with a two sigma fading BW of 1 Hz. Each condition was measured for 2 h.

The measured performance, employing the maximum interleaving period (the 72-frame “Very Long” interleaver), under each of the conditions listed for coded 1×10^{-4} BER is shown in Table 17.

TABLE 17

High data rate mode performance tests for 1×10^{-4} BER

User data rate (bit/s)	Average signal-to-noise ratio (dB) for BER not to exceed 10^{-4}		
	AWGN channel	Rician channel	Poor channel
12 800[(1)]	27	–	–
9 600	21	30	30
8 000	19	25	26
6 400	16	21	23
4 800	13	17	20
3 200	9	12	14

(1) Optional data rate.

1.6.4 Doppler shift performance

During Doppler shift performance test modem acquired and maintained synchronization for at least 5 min with a test signal having the following characteristics: 9 600 bit/s/very long interleaver, ±75 Hz frequency offset, 2 ms delay spread, a fading BW of 1 Hz, and an average signal-to-noise ratio of 30 dB.

1.7 Associated communications equipment

The QAM constellations described in this Annex are more sensitive to equipment variations than the PSK constellations described elsewhere in this Recommendation. Because of this sensitivity, radio filters will have a significant impact on the performance of modems implementing the high data rate waveform. In addition, because of the level sensitive nature of the QAM constellations, turn-on transients, AGC, and ALC can cause significant performance degradation.