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(54) A METHOD OF TRANSMITTING INFORMATION

(71) We, LICENTIA PATENT VERWALTUNGS G.M.B.H., of 1 Theodor-Stern-Kai, 6 Frankfurt/Main 70, Federal Republic of Germany, a German body corporate, do hereby declare the invention for which we pray that a patent may be granted to us and the method by which it is to be performed to be particularly described in and by the following statement:

The invention relates to a method of transmitting information particularly in the short-wave range.

With radio and wire-connected data transmission paths there are generally propagation times which considerably limit the transmission speed when using conventional telegraphy systems. For this reason a bit step speed of only 50 to 100 bits per second is usual in telex methods in the stop-start operation for example. In order to remedy this deficiency there has been a change to wire-connected data transmission paths in systems having so-called echo or propagation time equalizers. According to the knowledge of modern theoretical information technology the use of a propagation time (echo) equalizer guarantees optimization with respect to data transmission speed, transmission output and error rate in accordance with the parameters of transmission channels. While the echo equalizer has generally succeeded in the meantime on rapid wire-connected data networks (for example communication networks between computer centres) its use in radio networks has up till now been little known. This has several causes which will be stated in the following:

a) in the meantime transmission methods have been developed and introduced which however do not have the output data which may be achieved by systems having echo equalizers, but which could correspond to the output level required some years ago. All of the telegraphy methods which are characterized as multiple-tone narrow band systems, for example "Kathryn" or "link systems", belong to this group.

b) The parameters of the radio paths change very rapidly in general even in completely stationarily operated transmission/reception systems. This makes it necessary to have a constant adaptation process of the echo equalizer for matching to the instantaneously present operating parameters of the used "channel". As a result of current changes in the reflecting layers of the ionosphere repetition of the equalizer setting in 100 ms intervals is necessary for example in the short-wave range.

c) A further difficulty with regard to wire-connected data connections is presented by the propagation time distortions which are very much more pronounced in radio paths. These distortions are caused by multi-path broadening out of the signal energy irradiated by the transmitter and detected by the receiver.

With the Link-11 system for the short-wave ranger (2 to 30 MHz) a single sideband transmitter with linear amplifier final stages is used at the transmission end. The data signal is divided up in a modulator part connected in front of the transmitter into 15 independent narrow-band parallel channels each having a bandwidth of less than 100 Hz. Thus 4-phase shift keying (DPSK) is used with a transmission speed of approximately 80 Baud per individual channel. The overall bandwidth of all 15 parallel channels, the so-called "system bandwidth", amounts to approximately 2 kHz, since the individual frequencies (tones) have a spacing of approximately 130 Hz in each case. The "data tone spectrum" of the 15 individual channels is transmitted by the transmitter as a single sideband signal. Since all 15 tone frequencies are transmitted at the same time, the transmission output available is evenly divided over all tone frequencies. Because of the principle of linear superimposition however the summed output drops with the number of channels. This loss of output constitutes the main disadvantage of this multi-tone telegraphy method in conjunction with a further loss which is caused by the 4-phase shift keying in comparison to serial systems having 2-phase shift keying. A further disadvantage of these systems can be seen in its susceptibility to selective fading. With a double-path propagation having almost the same receiving field strength and a propagation time difference of 2 msec for example "gaps" in the transmission spectrum arise periodically at a spacing of about 500 Hz, i.e. with a 3 kHz wide spectrum, six fading frequencies are obtained which have the result that some of the 15 parallel channels supply erroneous transmission results. There is also an additional disadvantage that the simultaneous transmission of 30 bits at 15 tone frequencies makes linear transmission end stages necessary (single sideband transmitter).

From the publication "Proceedings of the IEEE, vol. 56, No. 10 (October 1968), pages 1653 to 1679 it is furthermore known to transmit a correlating test pulse sequence before the actual transmission of information, to determine from this the pulse response of the transmission channel at the receiving end by means of compression and to cause corresponding matching of the equalizer filter parameters. The disadvantageous part of this method is the fact that on the one hand it does not take the rapid changes in the pulse response of a short-wave transmission path into consideration and on the other hand — although the pulse compression and thus determination of the pulse response is erroneous when there is a deviation in frequency between the transmission and receiving end — it is possible not to take the determination and compensation of these frequency shifts into consideration.

The invention seeks to create a method of the type stated at the outset in which sufficiently accurate matching to the respective current pulse response of the transmission channel is ensured.

According to a first aspect of the invention, there is provided an information transmission method comprising transmitting the information as data frames containing a test sequence between a first and second information block, storing the received first information block while determining the current pulse response of the transmission channel using the test sequence of the associated data frame and adapting the parameters of an equalizer filter to this pulse response and thereafter equalizing the first and second data blocks using the adapted filter.

According to a second aspect of the invention, there is provided a method according to Claim 1, characterized in that the bits of the second information block incoming during equalization of the first information block are stored in intermediate manner.

In accordance with a preferred embodiment, the bits of the second information block incoming during equalization of the first information block may also be stored in an intermediate manner. A pause of several bits may be provided for decay of echoes between successive data frames and in each data frame, between the test sequence and each of the first and second information blocks. In a preferred refinement, the first and second information blocks may comprises 130 bits in each case 10 bits of each of these being redundant, the test sequence comprising 13 bits and the pauses between the data frames and the pauses between the test sequence and the first and second information block comprising 9 bits in each case. The information is transmitted largely at a speed of 3000 Baud. In preferred manner, an optionally correlated code may be used as the test sequence. The current pulse response of the transmission channel may be determined by compression of the respective test sequence. The equalized information blocks should be stored in intermediate manner and should be transmitted continuously frame by frame either serially or in parallel.

In accordance with a further refinement, the phase difference φ corresponding to the frequency deviation of the carrier, added on the receiver side, from the desired value may be determined and compensated at the reception end from the amplitudes of the compressed test sequences for every two successive data frames in a sine channel and a cosine channel orthogonal thereto in order to ensure accurate pulse compression and thus determination of the pulse response in a manner which is as free of errors as possible. Thus in order to compensate the phase difference φ, the signal of the sine channel and that of the cosine channel is multiplied on the one hand by the cosine of this phase difference φ and on the other hand the product of the signal of the cosine channel having the negative sine of the phase difference φ and the product of the signal of the sine channel having the sine of the phase difference φ is added to the thus multiplied signal of the sine or cosine channel.

The invention will now be described in greater detail, by way of example, with reference to the drawings, in which:—

Figure 1 is block diagram of two-transmission/reception stations for use with the method of the invention:

Figure 2 shows the time path of the data transmission:

Figure 3 is a block diagram of a demodulation arrangement:

Figure 4 shows the filter setting and procedures to be carried out in the processor of the equipment.

Figure 5 shows a list of the arrangement of figure 3 with an arrangement for compensating phase deviation.

Figure 6*a* shows a signal sequence sent out.

Figure 6*b* to 6*e* are diagrams for clarification of the relationship in the two orthogonal channels after pulse compression and,

Figures 7*a* to 7*d* shows an example of determining the rotation angle from two heavily rotating 13 bit Barker sequences.

A data transmission system having a propagation time equalizer adapated to the transmission path circumvents the difficulties described at the outset in conjunction with the known systems. In principle it is a question of distributing the information to be transmitted over the entire transmission band available, which in the short-wave range is 3 kHz wide for example, so that when losing a part of the spectrum there are still no or only insubstantial transmission errors. This means that the spectrum of the information to be transmitted must be as "white" as possible. The latter is achieved by an appropriate automatic coding of the information as well as a high telegraphy speed of transmission. Since the propagation time equalizing at the receiver end permits a telegraphy speed matched to the bandwidth of the system (in short-wave range with 3 kHz bandwidth up to 2400 Baud with binary coding) it is also possible to use the transmission bands available optimally. Because of the high telegraphy speed, the individual data bits are transmitted serially. Thus the transmission output available can be completely exhausted in each data bit. A further advantage resulting from serial transmission of data lies in the possibility of using less expensive and not necessarily linear transmitter end stages as are often used, for example in the VHF range in B-operation. This makes it possible to use fully transistorized power stages which are already available having an output of up to 2 kW with all of the known advantages with respect to linear valve final stages such as a considerably higher degree of efficiency, for example, a smaller weight and volume and simpler tuning etc.

The essence of the invention is to be seen in the fact that the receiving end equalizer is set in accordance with the high time variation characteristics of ionosphere radio paths in the short-wave range at regular time spacings of approximately 100 msec to the current pulse response of the transmission channel in each case. For this reason, in order to probe the transmission path, a short test sequence in the form of an optimally correlated bit sequence is transmitted at a time spacing of approximately 100 msec in each case, from which bit sequence the current pulse response of the transmission path in each case is determined at the receiving end by means of pulse compression and appropriate matching of the equalizer/filter parameter is undertaken. The actual information is transmitted between the individual test sequences in the form of data or information blocks whereby, for every two information blocks with a test sequence in between, one data frame is formed and the first information block of a data frame is stored in intermediate manner in order to reduce the "ageing" of the pulse response (this will be discussed in greater detail later) until the related test sequence is evaluated and until the equalizer/filter parameters is adapted. The important thing for perfect determination of the pulse response is determining and compensating frequency deviations between the transmission end (suppressed) carrier and the receiving end added carrier to which the invention also relates and which will be discussed in greater detail later.

Fig. 1 shows two transmission/reception stations in a block diagram for serial rapid telegraphy in the short-wave range these having adaptive echo equalizers. Each station is equipped with the normal transmitter/receiver (ESB means "single sideband"). A device designated as a "real, time processor" for modulating the signal to be transmitted (modulator) and for linear equalizing and demodulation of the received signal is connected between the transmitter/receiver and a data source and a data receiver.

In fact the method in accordance with the invention is suited for a so-called "Off-Line" operation (not real time), in which simple standard computers are sufficient for processing the signals, but the advantages of the method in accordance with the invention are important when using it with quasi-real time operation, for example of a digital speech transmission.

The type of modulation used must be linear. This is the compulsory result of the following marginal conditions:

The useful bit rate should reach 2400 Bd. Taking into account the high degree of use of the short-wave range, the standard bandwidth of a telephony channel of 3 kHz should not be exceeded. Since large linear signal distortion is to be expected on the transmission paths an echo equalizer must be used.

It is advisable if a specific type of vestigial side band modulation is used as the type of modulation, but this is not the subject of the present invention. The single disadvantage i.e. the distortion of the signal transmitted which is caused by this type of modulation proves on closer observation to be insubstantial since as a result of using the test sequences for receiving end measurement of the channel distortions, the adaptable equalizer can be set in completely directed manner so that it overcomes both the linear modulation distortions and the linear distortions caused by the short-wave transmission path.

Fig. 2 shows the time path of the data transmission. The digital information is transmitted in the form of data frames. Each data frame contains a test sequence for determining the channel parameters, a first information block being in front of and a second information block being behind the test sequence.

Energy pauses are provided between successive data frames and in the individual frames between the test sequence and the transmission blocks, in which pauses echoes are able to fade away. The first and second transmission block of a frame contain the same number of binary telegraphy steps in each case for the actual useful information as well as some redundance bits for error recognition at the receiving end. When using in the short-wave range for bridging distances up to 2000 km the following arrangement of the data frame shown in Fig. 2 is advisable. The transmission duration of the data frame including the energy pause after the preceeding data frame amounts to 100 msec. Since echoes are to be expected at propagation time differences up to approximately 2 msec, 3 msec are provided in each case for the energy pauses and this corresponds to 9 bits in each case with a telegraphy path speed of 3,000 Baud. Thus, in toto 9 msec or 27 bits are allocated to the three energy pauses required per data frame. The binary phase-shifted 13-bit Barker sequence is used as a test sequence, approximately 4 msec being necessary for its transmission. Since the total time not useful for the actual information transmission per data frame may amount to 20 msec, approximately 7 msec still remain, deducting the energy pauses and the test sequence, this being used for data security purposes, and in fact 40 msec or 120 bits in each case are allocated to the first and the second information block for the information to be transmitted and approximately 3.5 msec or 10 bits are used for redundance.

These 20 bits of redundance per data frame correspond to a Hamming distance of at least 4.

As a result of dividing the data frame into two, into a first and second information block, and as a result of the said intermediate storage of the first information block, an "ageing period" of the equalizer setting of about only 50 msec is achieved because of the fact that the test sequence is arranged between the first and second information block in the centre of the data frame, although the equalizer processor redetermines the weights or filter parameters of the equalizer/transversal filters shown in the form of computer algorithms actually in 100 msec spacings. In this way stability and accuracy are obtained — thinking only of the agreement between the (depressed) carrier on the transmission side and the added carrier on the receiving side which have to be kept stable in terms of frequency — however additional processing time in the processor must be taken into account since as a result of intermediate storage of the first information block until evaluation of the test sequence and setting of the equalizer weighting approximately 50 msec of additional throughput time in the processor must be added. One can assume that the computer time for setting the equalizer weighting is negligible with respect to the duration of processing of the actual information instruction. In consequence, the total throughput time of the quasi-real time equalizer processor is measured by the incidence of the beginning of the data frame until the first information bit of this frame is transmitted to the data receiver after 100 msec.

The reason for using test sequences for the equalizer setting — pulse response measurement is also possible with a single subpulse (elementary character of the linear telegraphy) — can be seen in the fact that on real radio paths such bad interference spacing ratios are often given that the measurement of the channel parameters must be carried out almost always with an improved signal interference spacing in order that the method of the invention will lead to the required result. The required interference spacing improvement when determining the pulse response is achieved by pulse compression of the test sequence. When using the 13-bit Barker sequence as a test sequence, the interference spacing gain with respect to the white noise amounts to more than 10 dB. The interference spacing of the pulse response measurement is thus at approximately 25 dB if one assumes a channel having only 15 dB useful signal/noise ratio. In practice, usable ionospheric connections from approximately 20 dB signal-to-noise ratio of the transmission channel are expected as compared to white noise. Since, at the moment in the short-wave range, transmitters are usually operated for slow telegraphy, the same channel interferences usually act as a continuous wave i.e. in the base band as an interference of constant frequency and amplitude. These interferences do not have any greater effect at least than white noise, for example at the same signal to noise ratio.

In summary therefore, it can be stated that the 13-bit Barker sequence is sufficient for the signal to noise ratios arising in practice.

Fig. 3 shows the arrangement for ("quadrature") demodulation of the transmitted and carrier signal at the receiving end (quadration is a non-linear procedure and thus cannot be used here). The arrangement corresponds to that of a coherent receiver. The equalization takes place in the base band. Two channels — a sine channel and a cosine channel — are necessary for "quadrature" demodulation in which the linearly distorted carrier receiver signal is mixed multiplicatively with a carrier signal prepared by a superimposition oscillator in one case directly and in another case after a phase shift of the carrier by 90°. Since two channels (the sine and the cosine channel) must be processed in the echo equalizer processor, the pulse response has to be determined in two channels as well. In order to be able to use the whole of the energy of the useful signal available in one channel, a filter, time-inverted with respect to each channel, a so-called matched filter M.F., is contained in the sine and in the cosine channel, said filter being constructed as a transversal filter and its weighting correspond to the time-invereted pulse response of each respective channel. As a result of this matched filter, non-linear quadration is avoided (the method should be linear). The matched filters are connected together at the output end in an adder the starting signal of which is supplied to a reciprocal filter R.F. The equalized data is available at the output of the reciprocal filter.

The arrangement in accordance with Fig. 3 has the following advantageous features:

—Determining the coefficients (weightings) of the matched filter is very simple.

—The signal energy in the channel is used optimally because of the additional use of the energy of the echo.

—The instantaneous phase of the carrier added during quadrature demodulation does not play any part, the matched filter replaces the (non-linear) quadration.

—The output signals of the matched filter are real, they can be added and then further processed in one channel.

—When using a vestigial side band modulation with small amplitude modulation, the output signals of the matched filters only have a small vestigial distortion.

—The reciprocal filter only needs to equalize the frequency response of the channel.

—The scanning time can be selected as desired if the scanning theorem is fulfilled.

—The reciprocal filter is set independently of the number of scanning processes per telegraphy step as if every telegraphy step was only scanned once. As a result, the reciprocal filter is substantially simplified since the number of its weighting depends still on the number and strength of the echoes or their configuration.

—The use of the matched filter procedure on the pulse response of the channel leads to a "symmetrical pulse response" which is decisive for setting the weightings of the reciprocal filter.

—The reciprocal filter represents a filter which is inverted to the symmetrical pulse response in the frequency range. In the time range the reciprocal filter is the equalizer filter of the symmetrical pulse response. The weightings of the reciprocal filter are set with the aid of a very rapidly converging iteration method in the time range.

Fig. 4 shows an overall view of the filter setting and procedures to be carried out in the processor. Test sequences and information blocks are processed separately and in fact filter settings are obtained in a so-called control parth from the test sequences, these settings being used in a so-called data path for the scanning values of the information blocks. The test sequences are fed, together with their echoes, into a sine and cosine channel of the control path initially via a compression filter optimized with respect to the pulse response measurement and designated as an optimal filter. It is a question of a firmly set transversal filter having a total of 29 weighting, the structure of which depends on the number of scanning values per telegraphy step. This optimal filter is the same in both channels, i.e. in the sine and in the cosine channel, this is only true for the 13-bit Barker sequence.

The transversal filters (matched filters) time-inverted and set from the pulse responses obtained by means of the optimal filters are designated "autocorrelation of the pulse response" in Fig. 4 in the control path and, in the data path which also has a sine and a cosine channel, are designated as "cross correlation of the information". In fact, the use of the matched filter procedure on the pulse response corresponds to an autocorrelation of the pulse response. The autocorrelation function of the pulse response designated farther above as "the symmetrical pulse response", always has a positive (largest) main value which is always in the middle. This main value contains the sum of all echo quadrates and thus the entire signal energy of the channel (by echoes we mean all signal values of the pulse response in question). since the cross-correlation filter is set identically in the sine channel of the data path as the matched filter in the sine channel of the control path and the cross-correlation filter is set in the cosine channel of the data path identically to the matched filter in the cosine channel of the control path, the information in the data path is cross-correlated with the matched filters of the pulse response of the respective channel. As a result of this cross-correlation, a main value is obtained for each data value (bit step) its appearance time being known, and this simplifies scanning. We have already discussed the reciprocal filter and the repetition method for setting it above. The data frames are equalized frame by frame. In addition the intermediate frequency output signals of ths high-frequency receiver are scanned, subjected to analog-digital conversion and are stored in the interim until the equalizer filters are set. The bytes (data values) stored in the interim are then demodulated by the signal processor and equalized by the adaptive equalizer filters. The equalized binary data values then reach an intermediate store again. An output procedures ensures that the equalized data values deliver frame by frame are continuously passed in a serial or parallel bit flow.

Since the test sequences are sent out only every 100 msec matching of the equalizer filters to the current pulse response of the transmission channel is only possible during these time spacings. Should the transmission channel change within 100 msec, then the adaptive filters are only correct at the beginning of the 100 msec interval, i.e. the filters "age." Considerable errors can be attributed to oscillations in the instantaneous phase and amplitude of the transmitted signal. Instability of effectively 1dB in the amplitude and effectively 7° in the phase involves an additional variance of –20 dB in binary phase shift keyed signals which has an effect like white noise at a 20dB signal to noise ratio. In this connection it is of interest as to how accurately the frequency of the carrier of the demodulation oscillator which is added at the receiver end to the reception signal has a to agree with the carrier frequency of the received signal, so that no further noticeable errors arise. If constant detuning of the demodulator frequency is stipulated by the carrier frequency when starting from a stable channel, then a permissible 10° phase deviation in the time span from the beginning to the end of an information block of a phase rotation of 20° per data frame (or 100 msec) of a phase rotation of 180° per second correspond to this which equals a frequency deviation of 0.5 Hz. This stability cannot however be kept even at transmission frequencies up to 15 MHz at a quartz stability of 10^{-7}.

For the purpose of establishing possibilities of compensating the effects of detuning between the carrier frequency of the received signal and the frequency of the demodulation oscillator, reference will be made in the following to the method of writing the Z-transformation. If M (z) is the transmitted information sequence, S (s) is the pulse response (scanning) sequence in the sine channel and C (z) is that in the cosine channel, thus, for the received sequence of scanning values E (z), the following is true while ignoring detuning:—

$$E(z) = M(z) \cdot C(z) + j\, S(z)$$

If after pulse response measurement and setting of the matched filters has taken place, there is a phase deviation φ as a result of the frequency deviation between the demodulator oscillator and the carrier of the received carrier signal, then the measured pulse response relevant for $\varphi = O$ is no longer current.

The new pulse response caused by phase rotation φ reads as follows:

a) in the cosine channel: $C(z).\cos\varphi - S(z).\sin\varphi$ (instead of C(z)),
b) in the sine channel: $S(z).\cos\varphi + C(z).\sin\varphi$ (instead of S(z)),
i.e. it is true that for the received sequence of scanning values, $E_c(z)$ is in the cosine channel and $E_s(z)$ in the sine channel:

$$E_c(z) = M(z) \,.\, [C(z) \,.\, \cos\varphi - S(z) \,.\, \sin\varphi]$$
$$E_s(z) = M(z) \,.\, [S(z) \,.\, \cos\varphi + C(z) \,.\, \sin\varphi]$$

If $\varphi = O$ i.e. if the said frequency deviation is equal to zero, which cannot be carried out in practice — as mentioned — then the error terms disappear and the following is true:

$$E_c(z) = M(z).C(z) \text{ and } E_s(z) = M(z).S(z).$$

In accordance with the present invention, the errors caused by a phase deviation φ can be compensated in accordance with the present invention by the following two measures.

1. Determining the frequency deviation at the moment of the matched filter procedure.
2. Use of an arrangement according to Fig. 5.

Measure 1 will be discussed later, but initially the arrangement according to Fig. 5 will be described in greater detail in the following.

Fig. 5 shows the part of the arrangement according to Fig. 3 following the quadrature demodulator complemented by an arrangement in accordance with the invention for compensating the phase deviation φ. The matched filters in the sine and cosine channel are designated S(1/z) and C(1/z) here, since they have the pulse responses S(1/z) or C(1/z) as filters inverted to their channel. The addition lies in the fact that the signal in the sine and in the cosine channel is multiplied by cos φ and then the product from the original signal of the cosine channel and $-\sin\varphi$ or the product of the original signal of the sine channel with $+\sin\varphi$ is added to the signal of the sine or cosine channel respectively. After adding the starting signals of two matched filters, the following starting sequence A(z) is obtained at the adder:

$$\begin{aligned} A(z) = M(z) \,.\, \{ & [C(z).\cos\varphi - S(z).\sin\varphi].\cos\varphi.C(1/z) + \\ + & [C(z).\cos\varphi - S(z).\sin\varphi].(-\sin\varphi).S(1/z) + \\ + & [S(z).\cos\varphi + C(z).\sin\varphi].\cos\varphi.S(1/z) + \\ + & [S(z).\cos\varphi + C(z).\sin\varphi].\sin\varphi.C(1/z)\} = \\ = M(z). \quad & [C(z).C(1z) + S(z).S(1/z)], \end{aligned}$$

i.e. the starting sequence is free of any error terms, it is identical to that in a phase deviation of zero. Since with known frequency deviation — determining which is the subject of measure 1 which still has to be discussed — the related current phase deviation φ can be continuously calculated (linear interpolation) by means of simple multiplication of the frequency deviation by the respective time which has elapsed since the pulse response measurement, facilitates the arrangement in accordance with Fig. 5 i.e compensation of the errors caused by phase deviations.

In the following, determining the frequency deviation will be described in greater detail.

Fig. 6*a* shows the signal sequence sent out with its alternation between test pulse sequences and information blocks. The time spacing between two successive test pulse sequences amounts to Δt. The test pulse sequences are all equal while the information blocks are different according to the information contained in them.

Fig. 6*b* to *e* are intended to clarify the relationships in the two orthogonal channels in the base band (sine and cosine channel) after pulse compression has taken place in the signal processor.

Fig. 6*c* or Fig. 6*e* shows the compressed test pulse sequences (the information blocks are left out here since they are processed in the data path) in the cosine or sine channel in the case where $\Omega_{ist} = \Omega_{soll}$, i.e. in the case of agreement of the signal carrier frequency and the superimposed oscillator frequency. It is clear that, in this case, successive compressed test pulse sequences (i.e. the pulse responses) always have the same amplitudes in both channels.

If, on the other hand, because of a frequency deviation $\Omega_{ist} \neq \Omega_{soll}$, then the phase relationship changes between the signal of the superimposition oscillator in the demodulator and the 1F signal continuously from which a "rotating" signal results in the base band and thus there is also "rotation" of the compressed test pulse sequences.

In Fig. 6*b* and 6*d*, the case where $\Omega_{ist} \neq \Omega_{soll}$ is shown and, in fact, the frequency deviation there should correspond to a 90°A rotation from test pulse sequence to test pulse sequence (i.e. in the time Δt). As can be clearly seen in this case, the amplitude changes considerably in both channels from one (compressed test pulse sequence to the next.

If the rotation angle between two test pulse sequences which are successive at a time space Δt is equal to θ, then the following is true for the frequency deviation Δf which is to be determined:

$$\Delta f = \frac{\theta}{360} \cdot \frac{1}{\Delta t} \quad (\mathrm{Hz}),$$

i.e. with the selected example with $\Theta = 90°$ the frequency deviation would amount to:

$$\Delta f = \frac{90}{360} \quad \frac{1}{100.10—3} \quad (\mathrm{Hz}) = 2.5\ (\mathrm{Hz})$$

at a time spacing Δt of 100ms.

determining the rotation angle θ necessary for determining the frequency deviation (Δt is predetermined) will be discussed later.

The results of pulse compression with "a rotating" channel lead to the following conclusions:

1. With "a rotating" channel the pulse compression incorporates errors (because of insufficient suppression of the subsidiary maxima as a result of the "rotation"), therefore the smallest possible "rotation" per test pulse sequence length is to be sought after for the pulse response measurement itself (for example a "rotation" of less than 10°).
2. Determining frequency deviations is also possible sufficiently accurately with linear distortion of the channel (echo formation) if the rotary angle θ between successive test sequences which is caused by the frequency deviation is smaller than 180°. If the rotation angle θ is larger than 180° then the measurement of the rotation angle is no longer unambiguous (since the arctan function in k. 180° with k = 0, 1, 2... is ambiguous).

The reason for using the correlation (also called optimal filtering or pulse compression) of test pulse sequences in order to determine the frequency deviation is to be seen in that this method is not susceptible to linear distortion also occurring in radio transmissions.

Fig. 7 shows an example of determining the rotation angle θ from two heavily rotating 13-bit Barker sequences. The double test sequence sent out is shown in Fig. 7*a* (base band). It comprises two 13 Barker bit sequences with a pause of 5 bits between the two sequences. The pause is necessary in order to avoid additional measurement errors when determining the frequency deviation in highly distorting media (5 echo intervals). By the magnitude T_T is meant the telegraphy step duration in sec.

Fig. 7*b* shows the double Barker sequence rotating towards the right after reception (before compression) in perspective view. The signal function not known in the sine and cosine (quadrature) channel arises herefrom by projection on to the sine or cosine channel plane indicated, whereby the sine channel plane lies perpendicular and the cosine channel plane lies parallel to the drawing plane. The rotation should be −12° per bit step. The starting signals of the optimal filters of both channels are shown in Fig. 7*c* and 7*d*. Two maxima can be clearly distinguished at a spacing of 18 telegraphy steps (the number 18 arises from 13 Barker bit sequence + 5 bit pause). The rotation angle θ amounts to −216° (18 bits at (−12°) per bit). As can be seen directly the rotation angle θ can be determined by means of the relationship

$$\theta = -\arctan \frac{A_1}{B_1} + \arctan \frac{A_2}{B_2} \mp k.\pi$$

with k = 0, 1, 2, . . . whereby A_1 and A_2 are the amplitudes of two successive compressed test pulse sequences in the sine channel and B_1 and B_2 are the corresponding amplitudes in the cosine channel.

The frequency deviation Δf is then calculated as

$$\Delta f = \frac{\theta}{3600} \cdot \frac{1}{18} \cdot \frac{1}{T_T} (\mathrm{Hz})$$

whereby the formation of the number 18 above is explained in the example according to Fig. 7, $\Delta t = 18 . T_T$).

The analytical approach shows that determining the frequency deviation is exactly possible even when linear distortion is present (channel echoes).

A frequency deviation with a rotation angle θ of −216° was simulated on a computer and gave an accuracy of better than 1% despite the large rotation.

It should also be pointed out particularly that a further great advantage of the method and the arrangement according to the invention lies in that, instead of optimally correlating sequences and filters any correlating sequence and any short correlating filter matching this can be used which produces a clear main value (maximum) per sequence during compression.

When using the arrangement according to Fig. 5, both information blocks lying between the two test sequences must be stored in intermediate manner since two successive test sequences in each case must be determined for compensation of a possible phase deviation of the rotation angle θ is determined. This produces a minimum throughput time in the equalizer processor of at least 100 msec when using a data frame in accordance with Fig. 2.

WHAT WE CLAIM IS:—

1. An information transmission method comprising transmitting the information as data frames containing a test sequence between a first and second information block, storing the received first information block while determining the current pulse response of the transmission channel using the test sequence of the associated data frame and adapting the parameters of an equalizer filter to this pulse response and thereafter equalizing the first and second data blocks using the adapted filter.

2. A method according to Claim 1, characterized in that the bits of the second information block incoming during equalization for the first information block are stored in intermediate manner.

3. A method according to Claim 1 or 2, characterized in that a pause is provided between successive data frames and, in each data frame, between the test sequence on the one hand and the first and second information blocks on the other hand.

4. A method according to any one of Claims 1 to 3, characterized in that the test sequence comprises an optimally correlating code, for example a 13 bit Barker code, the first and second data block comprise 130 bits in each case, and of this 10 bits of redundance and the pauses between successive data frames as well as between the test sequence and the first and second data block comprise 9 bits respectively.

5. A method according to any one of Claims 1 to 3, characterized in that an optimally correlated code is used as a test sequence, e.g. a Barker 13 code.

6. a method according to any one of Claims 1 to 5, characterized in that the current pulse response of the transmission channel is determined respectively by means of pulse compression of the test sequence.

7. A method according to any one of Claims 1 to 6, characterized in that the equalized information blocks are stored in intermediate manner and are transmitted continuously frame by frame either serially or in parallel.

8. A method according to any one of Claims 1 to 7, characterized in that the phase difference φ corresponding to the frequency deviation of the added carrier from the desired value is determined and compensated at the receiving end from the amplitudes of the compressed test sequence for every two successive data frames in a sine channel and a cosine orthogonal thereto.

9. A method according to Claim 8, characterized in that the signal of the sine and that of the cosine channel on the one hand is multiplied by the cosine of the determined phase difference cos φ in order to compensate the phase difference φ and on the other hand the product of the signal of the cosine channel having the negative sine of the phase difference $-\sin\varphi$ or the product of the signal of the sine channel with the sine of the phase difference sin φ is added to the signal of the sine or cosine channel thus multiplied.

10. An information transmission system comprising means for transmitting the information as data frames containing a test sequence between first and second information blocks, a store for intermediately storing the received first information block, means for determining the current pulse response of the transmission channel using the test sequence of the associated data frame, an equalizer filter for equalizing the first data block from the store and the second data block and means for adapting the parameters of the equalizing filter to the determined pulse response prior to equalizing the data blocks.

11. An information transmission method substantially as described herein with reference to the drawings.

12. An information transmission system substantially as described herein with reference to the drawings.

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5 SHEETS *This drawing is a reproduction of the Original on a reduced scale*

Sheet 1

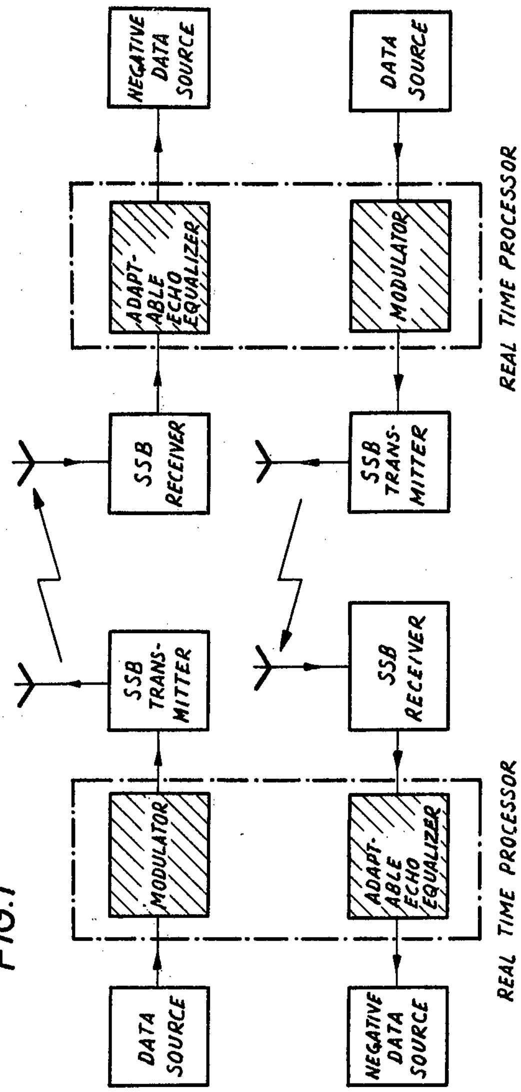


FIG.1

FIG. 2

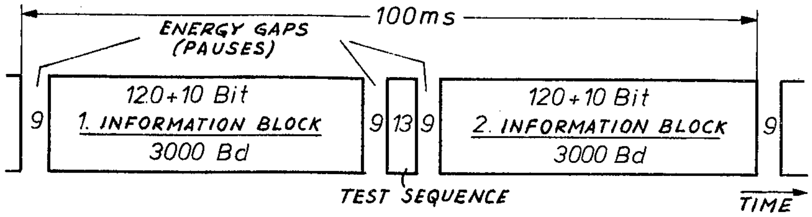


FIG. 3

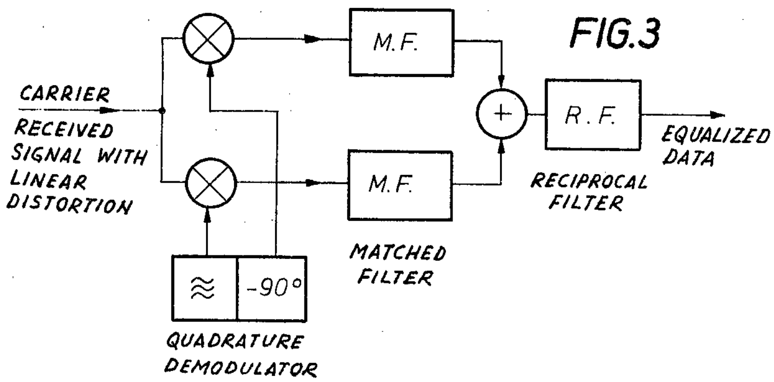


FIG. 5

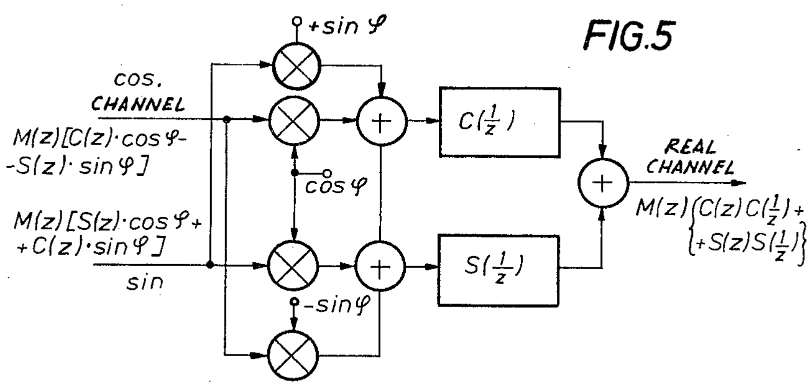
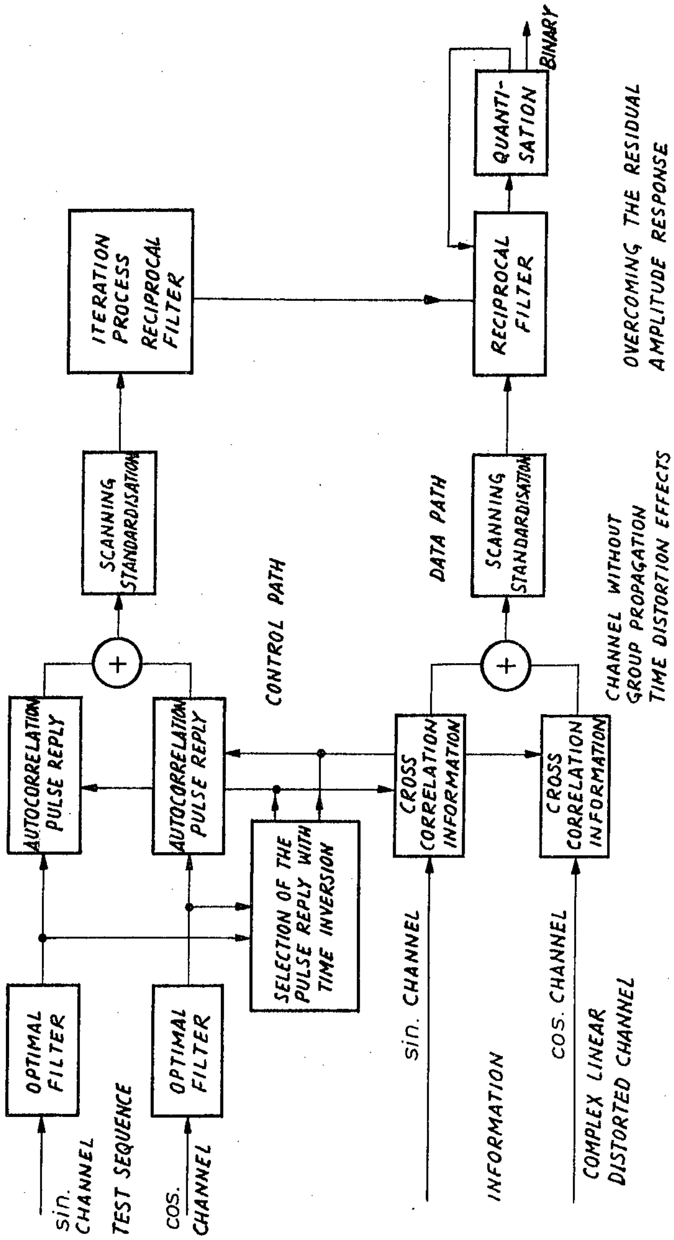


FIG. 4


sin. CHANNEL
TEST SEQUENCE
cos. CHANNEL
OPTIMAL FILTER
OPTIMAL FILTER
AUTOCORRELATION PULSE REPLY
AUTOCORRELATION PULSE REPLY
SELECTION OF THE PULSE REPLY WITH TIME INVERSION
SCANNING STANDARDISATION
ITERATION PROCESS RECIPROCAL FILTER
CONTROL PATH
DATA PATH
sin. CHANNEL
INFORMATION
cos. CHANNEL
COMPLEX LINEAR DISTORTED CHANNEL
CROSS CORRELATION INFORMATION
CROSS CORRELATION INFORMATION
SCANNING STANDARDISATION
CHANNEL WITHOUT GROUP PROPAGATION TIME DISTORTION EFFECTS
RECIPROCAL FILTER
QUANTI-SATION
BINARY
OVERCOMING THE RESIDUAL AMPLITUDE RESPONSE

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Sheet 4

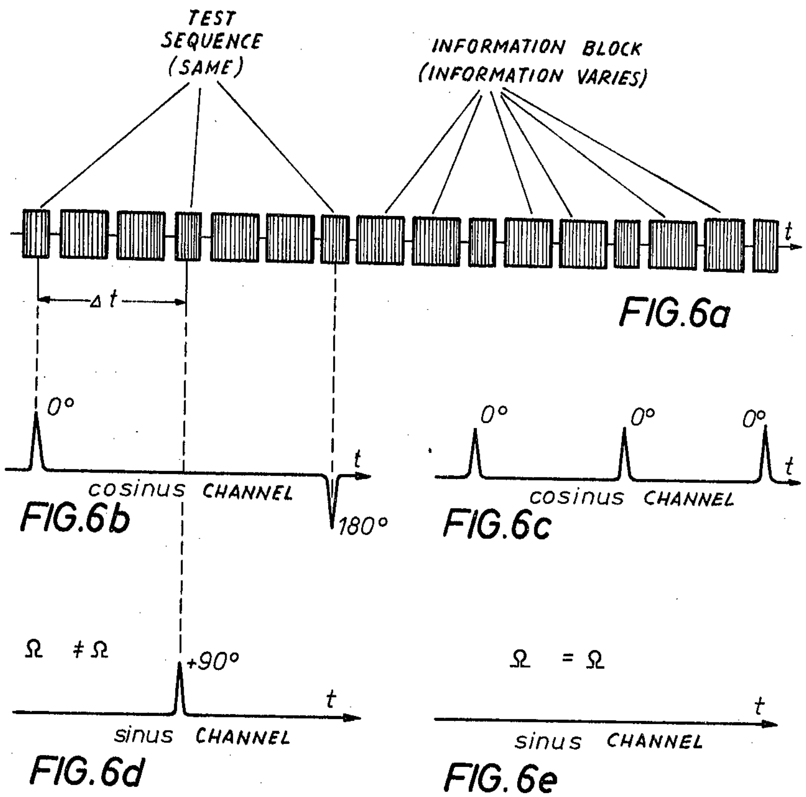

FIG.6a

FIG.6b

FIG.6c

FIG.6d

FIG.6e

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Sheet 5