

WIDEBAND HF SIGNALS DISPERSION DISTORTION COMPENSATOR BASED ON DIGITAL FILTER BANKS. THEORY AND APPROBATION

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The article proposes a computationally effective scheme of the dispersion distortions compensation device, based on the modification of the classical 2M-channel digital filter banks. The main idea of dispersion compensation is equalization of the delay and elimination of the phase shift, because the frequency dispersion of the signal is the delay of the signal at each frequency and the phase shift. The dispersion distortions compensation device involves the set of FIR filters and the delay line. The set of FIR filters eliminate fractional delay are implemented in the compensation scheme by means of the signal re-digitization with the help of the 19-th degree interpolation polynomial. Each of the FIR filters is multiplied by a coefficient that eliminates the certain phase shift. The delay line eliminates the integer part of the delay. Also, this article estimates the required number of channels in a dispersion compensation device based on a bank of digital filters and compares compensation algorithms on the basis of a filter bank and a filter-compensator. Analytical expressions are obtained for the approximate calculation of the maximum and standard deviations of the phase response of the system, reflecting the dependence of the phase response on the number of frequency subchannels, the order of the interpolation polynomial of the interpolator, the length of the polyphase components of the prototype filter, the slope of the dispersion characteristic, and the passband of the filter bank. Besides, a theoretical and experimental study is concluded, the purpose of which is to confirm the performance and efficiency (availability in terms of the signal-to-noise ratio) of the algorithms for compensating for dispersion distortion of broadband signals in the decameter band, embedded in a digital analysis-synthesis filter bank.

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Introduction

At present, decimeter radio communication is actively used for communication in remote and inaccessible regions of the world. Also, decimeter communication is an actual type of radio communication for regions that have suffered serious damage due to natural disasters.

Modern development of HF communication systems is aimed at increasing the information transmission rate by increasing the frequency band used [1-4]. When receiving signals with an extended spectrum, it becomes necessary to take into account the influence of the frequency dispersion of the ionospheric channel, since it leads to significant distortions of the signal and a decrease in the quality of information reception in general [5-7]. An actual problem is to assess the parameters of the frequency dispersion of the ionospheric channel and develop algorithms and devices for compensating for dispersion distortion of wide-band signals [8-10, 13-15].

Also, in broadband data transmission systems, digital filter banks have been actively used for signal processing for a long time [11, 12].

Filter banks are used to solve such problems as: estimating the power spectral density in radio monitoring tasks and cognitive radio systems, adaptive rejection of narrow-band interference, efficient message coding, etc. Therefore, for the decimeter range, it is relevant to use a filter bank and a dispersion distortion compensation device together. This problem is usually solved by using a special compensator filter (a filter consistent with the characteristics of the channel). However, in order to minimize the overall computational complexity of the algorithms for processing the received signals, it is urgent to introduce an algorithm for compensating dispersion distortions directly into the synthesis system of a bank of digital filters which are somehow used on the receiving side. Therefore, the compensation of dispersion distortions introduced into the bank of digital filters in order to eliminate the aforementioned losses is an urgent task.

Filter Bank for Dispersion Distortion Compensation

As a channel model with a dispersion frequency, a channel model with a linear dependence of the group delay of the signal on the central frequency (linear dispersion characteristic with slope s [3]) is adopted. The transfer function of the ionospheric channel in the absence of multipath signal propagation can be written in the form

$$\dot{U}(j2\pi f) = e^{-j\pi s f^2}, \quad f \in [-f_0/2; f_0/2] \quad (1)$$

where f_0 – is the sampling frequency, s – is the slope of the dispersion characteristic (inclination).

The phase response of the ionospheric channel is described by a quadratic frequency dependence, and the group propagation delay is described by a linear frequency dependence with the slope s $\mu\text{s}/\text{MHz}$.

Consider a system of banks of digital filters of analysis-synthesis with an ideal recovery. The bank consists of $2M = 64$ modulated complex exponent of analysis filters and the same number of synthesis filters [11,12,16]. Let us assume that $H_k(z)|_{z=e^{j2\pi fT}} = H_k(e^{j2\pi fT})$ and $F_k(z)|_{z=e^{j2\pi fT}} = F_k(e^{j2\pi fT})$ are respectively the frequency characteristics of the analysis filters and the synthesis filters, respectively, $T = 1/f_s$.

We can show that the signal distorted by the ionospheric channel at the output of the k -th analysis filter can be represented as

$$\tilde{x}_k(n) \approx x_k(n - n_k) e^{j\pi f_k^2 s}, \quad (2)$$

where $x_k(n)$ – the signal undistorted by the frequency dispersion at the output of the analysis filter, $n_k = \text{round}\left(\frac{f_k s}{T}\right) + \text{const}$ – the signal delay in the k -th channel of the analysis system, expressed in the sampling intervals T , $\text{round}(x)$ is the rounding function to the nearest whole.

The frequency dispersion of the signal is the signal delay at each frequency and the phase shift (see (2)), and for each frequency the delay and phase shift assume different values. Therefore, in order to eliminate the dispersion distortions, it is necessary to "equalize" the delay, which leads to the introduction into the bank's system of analysis and synthesis of an additional delay line before the synthesis filters and the $e^{-j\pi f_k^2 s}$ phase shift of the signals. Such a solution does not change the algorithm for calculating the synthesis bank filters in any way and does not depend on their specific type.

Figure 1 shows a computationally efficient compensation scheme for frequency dispersion of filter banks. To reduce the number of computational operations, the algorithm described in [12, 16] was applied. The main idea of such an algorithm is to represent synthesis filters (and analysis) in the form of a set of polyphase components of the original prototype filter $E_k(z)$, applying the FFT algorithm, and processing the signal at a low sampling frequency (i.e., before increasing M times).

In this case, it becomes necessary to introduce interpolation filters that compensate fractional delays.

In the proposed scheme, the compensation of group delay is carried out at a low sampling frequency. Moreover, one cannot limit oneself to introducing only compensation delay lines, since the delay compensation error will be times greater (since the sampling interval is times larger).

Therefore, a system of FIR filters $K_k(z)$, $k = 0 \dots 2M - 1$, eliminating fractional delay is implemented in the compensation circuit by means of the signal re-digitization with the help of the 19-th degree interpolation polynomial, and the delay line $z^{-\tilde{n}_k}$ eliminating the integer part of the delay, $\tilde{n}_k = \text{round}(f_k s M F_0)$, $\text{round}(x)$ – is a function that separates the integer part of the fractional number. Preliminary M_0 times interpolation is performed in order to reduce the signal distortion caused by the non-uniformity of the amplitude response and the nonlinearity of the phase response of the interpolating FIR filter. This allows, by processing one copy of the signal in the frequency domain, to reduce the amplitude-phase distortions that contribute to the interpolation FIR filter with group delay compensation. The value M_0 is chosen small: 1 (without pre-interpolation), 2 or 4. The length of the interpolation FIR filter is equal to $N_{\text{inter}} = N_{\text{poly}} M_0 - 1$.

The impulse response of the interpolation FIR filter is calculated as 19-degree Lagrange polynomial values at the corresponding time instants, provided that the input action, on the basis of which the polynomial coefficients are calculated, was unit-impulse. Moments of time correspond to the fractional delay value, which must be compensated.

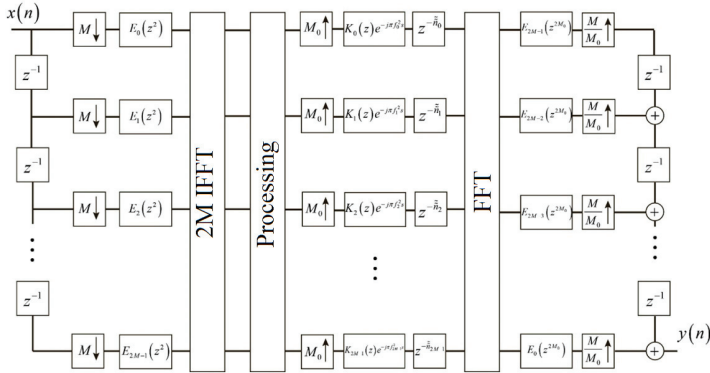


Figure 1. Computationally efficient filter bank with dispersion compensation

The transfer function of the channel-bank of the above filter bank is

$$T(z) = \sum_{k=0}^{2M-1} \sum_{l=0}^{M-1} U \left(e^{-j\frac{2\pi l}{M}} z \right) H_k \left(z e^{-j\frac{2\pi l}{M}} \right) K_k \left(z^{\frac{M}{M_0}} \right) e^{-j\pi f_k^2 z^{-\frac{M}{M_0}}} F_k(z)$$

where
$$H_k(z) = \sum_{q=0}^{2M-1} z^{-q} E_q(z^2) e^{j\frac{2\pi}{2M} kq} \quad \text{and}$$

$$F_k(z) = \sum_{u=0}^{2M-1} z^{-(2M-1-u)} E_{2M-1-u}(z^{2M}) e^{-j\frac{2\pi}{2M} ku} - c$$

putationally efficient representation of the transfer functions of analysis filters and synthesis filters, $E_q(z^{2M})$ и $E_{2M-1-u}(z^{2M})$ – polyphase components of analysis and synthesis filters.

The summation over q calculates the k -th coefficient of the inverse discrete Fourier transform (DFT) with the block length $2M$ (excluding the normalizing factor), where k is the filter index, and summing over u calculates the k -th direct DFT coefficient with the block length $2M$.

The figures Figure 3 and 4 show graphs:

1. Frequency response of the channel + bank with compensation system $T(z) \Big|_{z=e^{j2\pi fT}}$.
2. Frequency response of the aliasing components of the filter bank transfer function taking into account the channel and distortion compensation $T_{aliasing}(z) \Big|_{z=e^{j2\pi fT}}$.

The calculations were carried out for the bank of 64 filters ($2M = 64$) with the length of the filter $N = 1024$, satisfying the ideal reconstruction condition, in which aliasing components are mutually destroyed at the synthesis stage. It can be seen from the indicated dependencies that the superimposed components of the filter bank with compensation for dispersion distortions are not fully compensated, and as a result there is no complete signal recovery at the output of the filter bank.

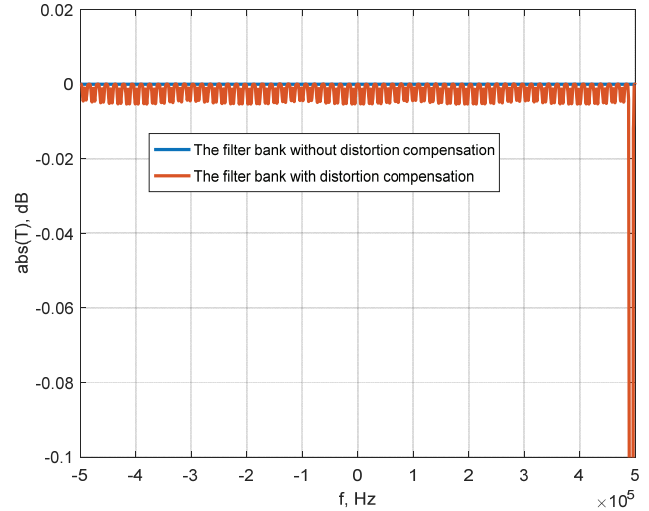


Figure 2. Frequency response of the channel + bank with compensation system

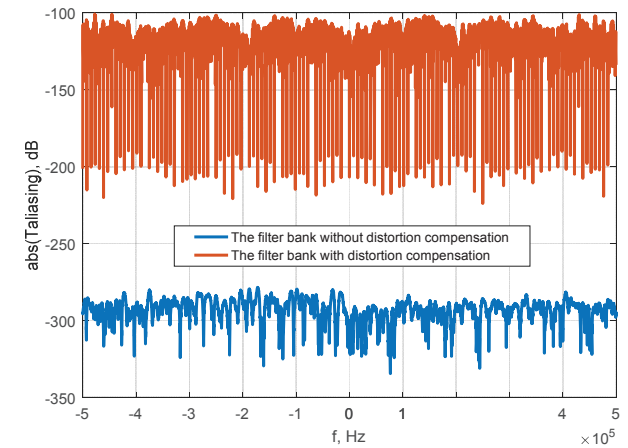


Figure 3. Frequency response of the aliasing components of the filter bank transfer function taking into account the channel and distortion compensation

The estimation of the minimum number of filters in the filter bank to ensure a given level of phase distortion

We can show [17] that the phase response of the useful component of the filter bank without taking into account the channel and without the compensation system is strictly linear and corresponds to the signal delay during processing:

$$\varphi_T(e^{j\hat{\omega}}) = \hat{\omega}(1 - N_{fit}) \quad (3)$$

Then, with compensation, the following condition must be met

$$\varphi_{T_0}(e^{j\hat{\omega}}) \rightarrow \varphi_T(e^{j\hat{\omega}}) + const. \quad (4)$$

In other words, the compensation should eliminate the channel effect, which is reflected in the appearance of the quadratic part in the phase response. For comparison, it is convenient to use the phase response of the channel + filter bank system after eliminating the linear part. In view of the above, (4) can be rewritten as

$$\tilde{\varphi}_{T_0}(e^{j\hat{\omega}}) = \varphi_{T_0}(e^{j\hat{\omega}}) - \hat{\omega}(1 - N_{fit}) \rightarrow const, \quad (5)$$

where $\tilde{\varphi}_{T_0}(e^{j\hat{\omega}})$ – the channel phase response + bank filter system after eliminating the linear part.

Figure 4 shows a graph of the phase response of the channel + system of the useful component of the filter bank transfer function with compensation at $s = 80 \mu\text{s} / \text{MHz}$ for a different number of filters after eliminating the linear part corresponding to the signal delay in the filter bank. The figure shows that the more filters are used, the better the quality of compensation (The channel phase response closer to zero, see the formula (5)). The filter near the frequency $\hat{\omega} = \pm\pi$ is not taken into account ($k \neq M$), because leads to a failure in the frequency response at the boundary with one of the neighboring filters, i.e. introduces distortions in the final frequency response of the compensator.

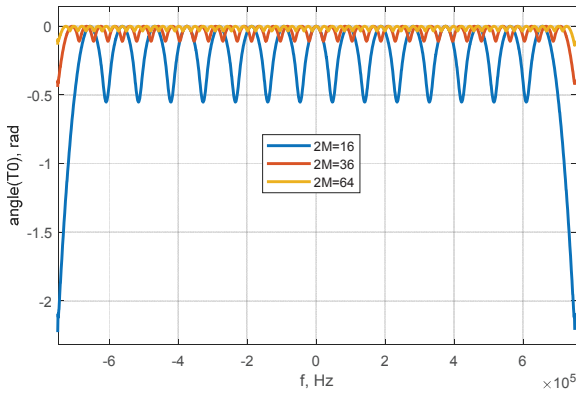


Figure 4. The phase response of the channel system + the useful component of the transfer function of the filter bank after eliminating the linear part for $s = 80 \mu\text{s} / \text{MHz}$

The necessary number of filters to ensure a given level of phase response deviation can be obtained from the solution of the optimization problem using the root mean square criterion (RMS) or minimax criterion. For both criteria, when solving the optimization problem, the variable parameter is the number of filters $2M$.

For a filter bank, the RMS criterion of optimality has the form

$$\varphi_{rms} = \sqrt{\int_F (\varphi(2\pi f) - \langle \varphi(2\pi f) \rangle)^2 df} \rightarrow \min_{2M},$$

$$\langle \varphi(2\pi f) \rangle = \int_F \varphi(2\pi f) df = const \quad (6)$$

where φ_{rms} – permissible deviation of the phase response from zero, F – frequency range (approximation interval).

The level of quality for minimax criterion is defined as the maximum deviation from the average between the maximum and minimum deviations from zero in the passband, because this average shows general phase shift of the signal and does not distort it. The minimax criterion for a filter bank can be written as

$$\varphi_{\min \max} = \frac{\max_{f \in F} (\varphi(2\pi f)) - \min_{f \in F} (\varphi(2\pi f))}{2} \rightarrow \min_{2M}, \quad (7)$$

where $\varphi_{\min \max}$ – permissible deviation of the phase response from average.

It is obvious that the phase deviation from zero is due to residual dispersion distortions of the signal within the passband of individual filters in the filter bank. The more filters there are, the narrower their passband, the less the amount of residual distortion.

We introduce the function

$$F(M, s, F_s) = \frac{\Delta f}{\Delta f_{coh}} = \frac{F_s}{2M} \sqrt{\frac{\pi |s|}{4}} = p_f, \quad (8)$$

where $\Delta f = F_s / 2M$ – bandwidth of one filter from the filter bank, p_f – dispersion coefficient of the channel formed by a separate filter from the filter bank, $\Delta f_{coh} = \sqrt{4 / (\pi |s|)}$ – channel coherence band with frequency dispersion.

Find $F(M, s, F_s)$ in such a way that

$$F(M, s, F_s) \cong const, \quad \varphi_{\min \max} = const \quad (9)$$

for the considered values of M, s, F_s and $\varphi_{\min \max}$.

Then from relation (3), $2M$ can be expressed for a known slope of the dispersion characteristic s , sampling frequency F_s , and a given level $\varphi_{\min \max}$.

From Figure 4 we can conclude that the phase deviation increases from the central frequency of the passband of a single filter and grows to its boundary. Given the channel model

$$\varphi_{\min \max} \sim \pi s \left(\frac{\Delta f}{2} \right)^2. \quad (10)$$

We seek (3) taken into account as

$$\sqrt{\varphi_{\min \max}} = C_1 F(M, s, F_s) = C_1 \frac{\Delta f}{\Delta f_{coh}}, \quad (11)$$

C_1 – constant.

As a result of numerical calculations for $M_0 = 2, N_{poly} = 16$ the sets of individual values $\varphi_{\min \max}$ and $F(M, s, F_s)$ the desired constant C_1 was found by the least squares method and the expression was found

$$\varphi_{\min \max} = 0.7073^2 F^2(M, s, F_s) = \left(\frac{0.7073 F_s}{2M} \right)^2 \frac{\pi |s|}{4},$$

$$C_1 = 0.7073. \quad (12)$$

Similarly, an expression was obtained for the standard deviation of the phase φ_{rms}

$$\varphi_{rms} = 0.575^2 F^2(M, s, F_s) = \left(\frac{0.575 F_s}{2M} \right)^2 \frac{\pi |s|}{4},$$

$$C_1 = 0.575 \quad (13)$$

Figure 5 shows a comparison of the theoretical (calculated by formulas (7, 8) and experimental dependences of φ_{rms} and $\varphi_{\min \max}$ on the number of filters $2M$ for four different values of the slope s and for the sampling frequency $F_s = 1.5 \text{ MHz}$.

Figure 5 shows that the larger the filter bank (the larger the parameter $2M$), the smaller the values of φ_{rms} and $\varphi_{\min \max}$ can

be obtained for different slopes of the dispersion characteristic. Also, note that the larger s , the more difficult the compensation, and accordingly, a larger filter bank is required to achieve the required values of φ_{rms} and $\varphi_{\min\max}$.

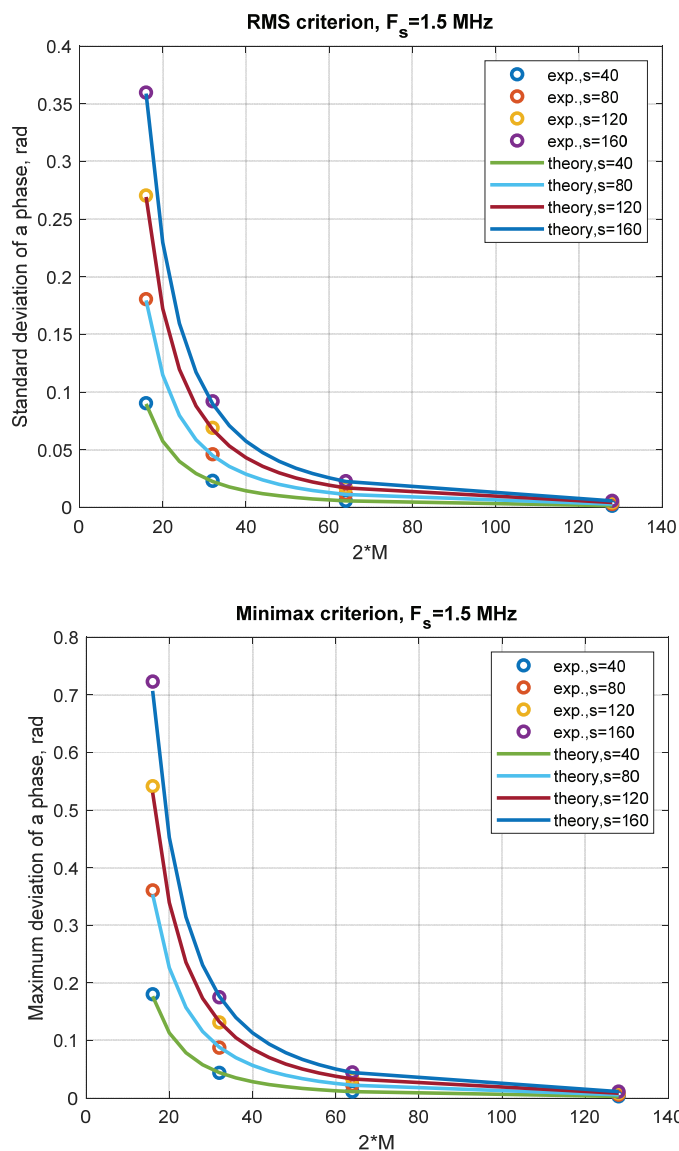


Figure 5. Theoretical and experimental dependences of the phase deviation on the number of filters in the filter bank, the RMS criterion and minimax criterion

Figure 6 shows the dependence of the number of arithmetic operations on the length of the interpolation polynomial of the interpolator filter N_{poly} for different values of M_0 and for $m = 8$, $2M = 64$. The figure shows that the number of arithmetic operations increases linearly with increasing N_{poly} . Also, with a 2-fold increase in M_0 , the number of computational operations also increases by about 2-fold.

Figure 7 shows the dependence of the maximum phase deviation on N_{poly} for various values of M_0 and for $m = 8$, $2M = 64$. The figure shows that the dependence of the maxi-

imum phase deviation on N_{poly} for various values of M_0 is small (maximum difference between maximum phase deviation for $M_0 = 1$ and $M_0 = 4$ is $0.3 \cdot 10^{-3}$ for $N_{poly} = 10$). Therefore, it is possible to set $M_0 = 1$ in the compensator in order to significantly reduce computational costs.

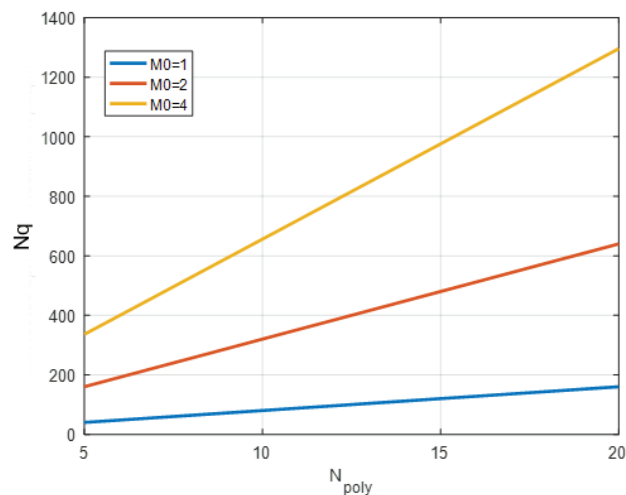


Figure 6. The dependence of the number of arithmetic operations on the length of the interpolation polynomial of the interpolator filter N_{poly}

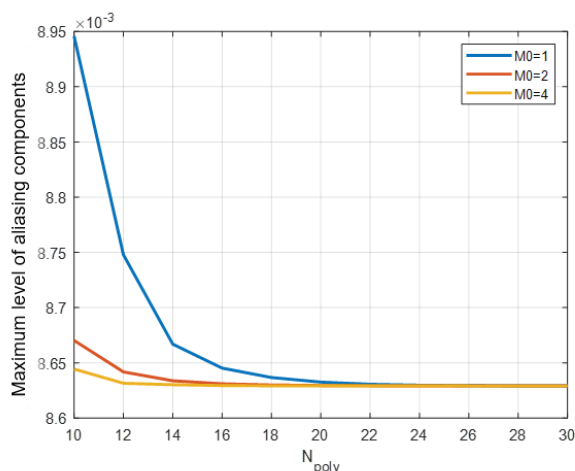


Figure 7. The dependence of the maximum phase deviation on N_{poly} for various values of M_0

The comparison of the computational efficiency of compensation algorithms in the filter bank

This section presents the results of comparing the computational efficiency of the proposed compensation device and filter-compensator. The filter-compensator is a matched filter (MF), consistent with the characteristics of the channel.

Tables Table 1 and 2 show a comparison of the number of real multiplication-addition operations in the compensator necessary to achieve the set values φ_{rms} and $\varphi_{\min\max}$ for different slopes s , $\mu\text{s} / \text{MHz}$ for $F_s = 1 \text{ MГц}$, $M_0 = 1$. Since the filter

bank implemented according to the computational scheme has the number of channels determined through 2^n , where $n = 2, 3, \dots$, then the levels φ_{rms} and $\varphi_{\min \max}$ can vary greatly change depending on $2M$. Therefore, when comparing fixed levels φ_{rms} and $\varphi_{\min \max}$, if the deviation in the filter bank reaches a lower value than the fixed one, this is indicated in brackets.

Table 1

The comparison of the number of material operations of multiplication and addition, which the filter compensator performs and the filter bank when compensating to achieve the specified values φ_{rms}

s , $\mu\text{s} / \text{MHz}$	φ_{rms} , rad	The number of channels in the filter bank, the length of the interpolation polynomial, and the number of real computational operations of the compensator $2M, N_{poly}, N_q$	The number of real computational operations of the filter-compensator
40	0.004	64, 10, 80	119
80	0.008	64, 10, 80	119
120	0.013	64, 10, 80	119
160	0. (for filter bank - 0.015)	64, 10, 80	118

Table 2

The comparison of the number of material operations of multiplication and addition, which the filter compensator performs and the filter bank when compensating to achieve the specified values $\varphi_{\min \max}$

s , $\mu\text{s} / \text{MHz}$	$\varphi_{\min \max}$, rad	The number of channels in the filter bank, the length of the interpolation polynomial, and the number of real computational operations of the compensator $2M, N_{poly}, N_q$	The number of real computational operations of the filter-compensator
40	0.015	32, 10, 80	166
80	0.01 (for filter bank - 0.0076)	64, 12, 96	182
120	0 (for filter bank - 0.012)	64, 12, 96	182
160	0.015	64, 12, 96	199

From tables 1, 2 we can draw the following conclusions:

1. The compensator in the filter bank with the same number of channels and computational operations provides approximately the same values of φ_{rms} and $\varphi_{\min \max}$, while the compensator filter, due to the large non-uniformity of the phase response, the values of φ_{rms} and $\varphi_{\min \max}$ can differ by an order of magnitude with the same number of computational operations;

2. The compensator in the filter bank for the given values of φ_{rms} requires computational operations less than 1.5 times than the filter-compensator, and for the given values of $\varphi_{\min \max}$, it is less than about 2 times.

The comparison of theoretical and experimental compensation results

The signal-to-noise ratio (SNR) at the MF output at the time the signal ends for a computationally efficient filter bank circuit with frequency dispersion compensation is defined as

$$\left(\frac{P_c}{P_u}\right)_{\text{эф. комп.}} = \frac{1}{N_{DFT}} \frac{\left| \sum_{i=-\frac{N_{DFT}}{2}}^{\frac{N_{DFT}}{2}-1} \dot{Q}_i e^{j2\pi i n_s} \right|^2}{2\sigma_u^2 \left(\sum_{i=-\frac{N_{DFT}}{2}}^{\frac{N_{DFT}}{2}-1} |J_i|^2 \right)}, \quad (14)$$

where

$$\begin{aligned} \dot{J}_i &= \dot{K}_{c\phi i} e^{j2\pi i n_s} \sum_{k=0}^{2M-1} \dot{H}_{ki} \dot{F}_{ki} \hat{K}_{ki} e^{-j\pi \left(\frac{f_k}{N_{DFT}} - i\right) s} e^{-j2\pi i \tilde{n}_k} + \\ &+ \sum_{l=1}^{M-1} \dot{K}_{c\phi(i+bl)} e^{j2\pi (i+bl) n_s} \sum_{k=0}^{2M-1} \dot{H}_{ki} \dot{F}_{k(i+bl)} e^{-j\pi \left(\frac{f_k}{N_{DFT}} - (i+bl)\right) s} e^{-j2\pi (i+bl) \tilde{n}_k}, \\ \dot{Q}_{sdi} &= \dot{S}_i \dot{U}_i \dot{K}_{c\phi i} \sum_{k=0}^{2M-1} \dot{H}_{ki} \dot{F}_{ki} \hat{K}_{ki} e^{-j\pi \left(\frac{f_k}{N_{DFT}} - i\right) s} e^{-j2\pi i \tilde{n}_k} + \\ &+ \dot{K}_{c\phi i} \sum_{l=1}^{M-1} \dot{S}_{i-bl} \dot{U}_{i-bl} \sum_{k=0}^{2M-1} \dot{H}_{k(i-bl)} \dot{F}_{ki} \hat{K}_{ki} e^{-j\pi \left(\frac{f_k}{N_{DFT}} - i\right) s} e^{-j2\pi i \tilde{n}_k} \end{aligned}$$

$\dot{S}_i, \dot{U}_i, \dot{H}_{ki}, \dot{F}_{ki}$ – DFT coefficients of the useful signal, impulse response of the channel, impulse response of the analysis and synthesis filters, respectively, \hat{K}_{ki} – the DFT coefficients of the impulse response of the filter-compensator fractional delay interpolated M/M_0 times M/M_0 , $\dot{K}_{c\phi i}$ – DFT coefficients of the impulse response of a filter matched to a useful signal $\dot{s}(n)$, $n = 0 \div N-1$, $b = \frac{N_{DFT}}{M}$ – proportionality coefficient. The number of DFT points N_{DFT} was chosen for reasons that a cyclic shift of the spectrum by $\frac{2\pi}{M}$ at normalized frequencies is equivalent to a cyclic shift of the DFT coefficients in the frequency domain by b .

Also, this section presents the results of an experiment to compensate for dispersion distortion in a wideband signal on a single-hop HF radiolink with a length of ~ 2900 km. The experiment used a wideband signal with direct spreading of the spectrum with a pseudo-random sequence consisting of 512 symbol and with a rectangular envelope function, the width of the main lobe of the signal spectrum was 400 kHz.

Among the experimental recordings of the signals were selected those where there is a minimum amount of station interference.

The operating frequencies were chosen near the MUF, 0.85-1.05 of the predicted one, since on them, the dispersion distortion is significant. Compensation was carried out by the methods proposed in the article, based on banks of analysis-synthesis filters. The dispersion characteristic slope was tentatively estimated by the maximum likelihood method.

In the course of the experiment, five different records were processed, the MF responses were constructed for the signal not processed by the bank, for the signal not processed by the filter-compensator and the compensation signal embedded in the computationally efficient filter bank scheme and also determined the gain in compensation (Figure 8).

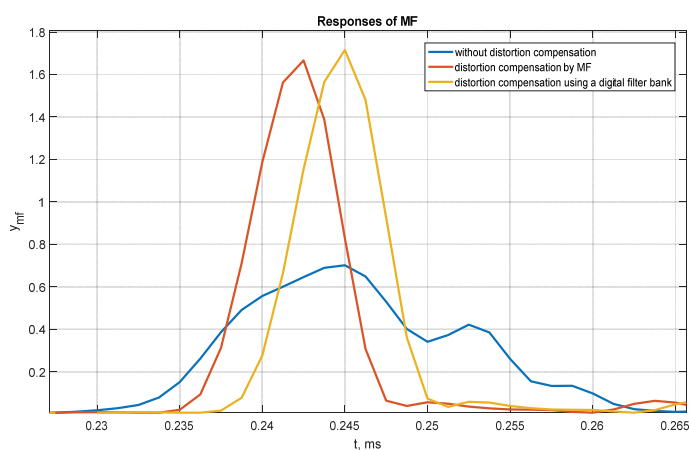


Figure 8. Experimentally obtained MF responses at $s = 140 \mu\text{s} / \text{MHz}$ (Record 1)

Compensation gain is determined by

$$K_{MF} = 20 \lg \left[\frac{\max(|\dot{q}_{MF}(n_*)|)}{\max(|\dot{q}(n_*)|)} \right] \text{ for the filter-compensator and}$$

$$K_{comp.bank} = 20 \lg \left[\frac{\max(|\dot{q}_{comp.bank}(n_*)|)}{\max(|\dot{q}(n_*)|)} \right] \text{ for computationally}$$

efficient filter bank scheme, respectively, $\dot{q}(n_*)$ – the response of MF without compensation, $\dot{q}_{MF}(n_*)$ – the response of MF with compensation by the filter-compensator, $\dot{q}_{comp.bank}(n_*)$ – the response of MF with compensation by the filter bank. The results of the experiment are summarized in Table 3.

The theoretical values of the gain in terms of SNR for different values are summarized in Table 4. The theoretical gain in compensation was determined as the difference between the SNR without compensation and with compensation. The theoretical value of the SNR is calculated by the formula (9).

Table 3

The results of experimental compensation for dispersion distortions

Record #	$s, \mu\text{s} / \text{MHz}$	K_{MF}, dB	$K_{comp.bank}, \text{dB}$
1	14	3.8	3.9
2	8	2.7	2.9
3	8	2.2	2.4
4	8	3.3	3.2
5	13	1.4	1.6

Table 4

The gain in theoretical compensation for dispersion distortion for a signal with a rectangular envelope function

$s, \mu\text{s} / \text{MHz}$	$K_{comp.bank}, \text{dB}$	K_{MF}, dB
80	4.	4.37
130	6	6.38
140	6	6.67

Conclusion

The article proposes the dispersion distortion compensator based on a bank of digital filters with a computationally efficient implementation. By calculation and using simulation, the operability of the proposed device was demonstrated. The expression was obtained for determining the signal-to-noise ratio at the output of the MF and the filter bank with compensation. In this article, we estimated the required number of channels in a dispersion compensation device based on a bank of digital filters. New expressions were obtained that allow, for a given level of the mean square or maximum deviation of the phase response of the channel + filter bank with compensation system, to determine the minimum required number of filters $2M$. Comparison of compensation algorithms was carried out on the basis of a filter bank and a filter-compensator, according to the results of which it was concluded that the compensator in the filter bank for the given values of φ_{rms} requires computational operations less than 1.5 times than the filter-compensator, and for the given values of φ_{minmax} , it is less than about 2 times. In addition, the article presents the results of field tests to compensate for dispersion distortions in a broadband signal and their comparison with theoretical data. The maximum gain in experimental compensation is 3.9 dB, and the minimum is 1.6 dB for various records. The theoretical gain is 4.36 dB for $s = 80 \mu\text{s} / \text{MHz}$, 6.37 dB for $s = 130 \mu\text{s} / \text{MHz}$ and 6.66 dB for $s = 140 \mu\text{s} / \text{MHz}$.

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КОМПЕНСАТОР ДИСПЕРСИОННЫХ ИСКАЖЕНИЙ ШИРОКОПОЛОСНЫХ СИГНАЛОВ ДЕКАМЕТРОВОГО ДИАПАЗОНА, ПОСТРОЕННЫЙ НА БАЗЕ БАНКА ЦИФРОВЫХ ФИЛЬТРОВ. ТЕОРИЯ И ЭКСПЕРИМЕНТ

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Аннотация

Предложена вычислительно эффективная схема устройства компенсации дисперсионных искажений, основанная на модификации классической схемы 2М-канального цифрового банка фильтров. Основанная идея компенсации дисперсионных искажений состоит в выравнивании временной задержки и устранении фазового сдвига сигнала, так как частотная дисперсия влияет на сигнал, внося дополнительную групповую задержку и фазовый сдвиг. Устройство компенсации дисперсионных искажений состоит из набора КИХ-фильтров интерполяторов и линии задержки. Набор КИХ-фильтров, устраняющих дробную задержку, реализован в схеме компенсации путем повторной оцифровки сигнала с помощью интерполяционного полинома Лагранжа 19-й степени. Каждый из КИХ-фильтров умножается на коэффициент, который устраняет определенный фазовый сдвиг. Линия задержки необходима для устранения целой части задержки. Также, предложен метод оценки необходимого количества каналов в устройстве компенсации дисперсионных искажений для достижения заданного качества компенсации, количественно определяемого среднеквадратичным и максимальным отклонением ФЧХ системы канал+компенсатор. Предложенный алгоритм компенсации сравнивается с алгоритмом компенсации на основе классического фильтра-компенсатора по качеству и сложности реализации. Получены аналитические выражения для приближенного расчета максимального и среднеквадратичного отклонения ФЧХ системы канал+компенсатор, отражающие зависимость отклонения ФЧХ от числа частотных подканалов, порядка интерполяционного полинома интерполятора, длины полифазных компонент фильтра прототипа, наклона дисперсионной характеристики и полосы пропускания банка фильтров. Представлено теоретическое и экспериментальное исследование, цель которого состоит в том, чтобы подтвердить работоспособность и эффективность (с точки зрения отношения сигнал/шум) алгоритма компенсации дисперсионных искажений широкополосных сигналов в декаметровом диапазоне, встроенного в банк фильтров цифрового анализа-синтеза.

Ключевые слова: декаметровая радиосвязь, широкополосные сигналы, дисперсионные искажения, банк цифровых фильтров, ионосфера.

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