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CLUTTER REJECTION IN A PASSIVE RADAR RECEIVER OF OFDM SIGNALS

The present invention relates to a passive radar
5 receiver receiving a radio signal comprising frames of
symbols each emitted on orthogonal coded carriers.

In the radar field, it is generally difficult, if
not impossible, to achieve the theoretically achievable
10 detection performance limits for a mobile target. This is
because detection performances are conditioned by the
signal to thermal noise ratio at the output of a tuned
filter in the radar receiver and in practise is generally
limited not by thermal noise but by clutter at the output
15 of the tuned filter. In the present context the term
clutter is to be interpreted in the wide sense of all
paths with zero Doppler effect. For example, for a
bistatic radar with distant emitter and receiver, the
clutter designates all of the following paths: the direct
20 path from the emitter and each path received following
reflection by a fixed obstacle.

Various methods of rejecting these unwanted signals
are known in the art but have nonnegligible drawbacks.
25 For example, adaptive rejection methods based on using a
covariance matrix of the signals received by an array of
sensors have the following limitations:

They eliminate only a limited number of
decorrelated interference signals, characterized by their
30 direction. Consequently, these methods are not optimized



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in the context of the fight against clutter when the latter is rich in multiple paths with different time-delays.

- They lead to the creation of blind axes, associated with the rejected signals, on which it becomes impossible to detect a target.

- They reject only signals whose signal to noise ratio is positive after angular compression. This rejection is limiting if it is effected at the beginning of radar processing, i.e. before distance-Doppler compression.

The invention is more particularly directed to the rejection of all zero Doppler effect paths in the clutter picked up by a passive radar receiver of particular orthogonal frequency division multiplex (OFDM) signals. OFDM signals are characterized by simultaneously emitting a large number of orthogonal sub-carriers phase coded with plural phase states or amplitude coded, i.e. by a spectrum of orthogonal lines, in the Fourier transform sense, over a finite duration T , equidistant at intervals $1/T$.

A bistatic radar disclosed in the French patent application FR 2776438 processes coded OFDM (COFDM) digital radio signals in the context of radio and television broadcasts conforming to the European Digital Audio Broadcasting (DAB) and Digital Video Broadcasting (DVB) standards. These signals, which are therefore sent by emitters of opportunity in the case of passive radar receiver applications, ensure optimum use of the spectrum

emitted, in a similar manner to white noise, and are resistant to multipath propagation and interference.

According to the above patent, the radar receiver comprises a plurality of receive antennas for detecting
5 the signals. The radar processing is based on Doppler-distance correlation of the signals received with a emitted signal time reference. The time reference is obtained by decoding the signals recorded conforming to the radio telecommunications operations effected.

10 However, because of the bistatic nature of the radar system, the power of the direct path signal is high compared to that of the wanted signal reflected by a target. The direct path should be rejected before effecting the Doppler-distance correlation. The energy
15 contained in the distance-Doppler secondary lobes of the direct path is generally significantly higher than the thermal noise, so that targets situated in the vicinity of the direct path are difficult to detect.

20 An object of the invention is to reduce or even to eliminate the contribution of the direct path and more generally of unwanted zero Doppler effect signals to the processing of receive signals prior to the Doppler-distance correlation.

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In order to reach this object, a radar receiver according to the invention, processing a radio signal received via a propagation channel and comprising frames of symbols each emitted on coded orthogonal carriers,
30 comprising shaping means for converting the received

signal into a digital symbol signal and Doppler-distance correlation means for discriminating mobile targets, is characterized in that it comprises filtering means for eliminating in the symbol signal at least unwanted zero
5 Doppler effect signals in order to apply a filtered signal including essentially signals backscattered by targets to the correlation means.

In a first embodiment, in particular for a radar
10 having only one receive channel, the filtering means comprises means for producing spectral lines of the symbol signal corresponding to the orthogonal carriers, means for detecting the unwanted zero Doppler effect signals by estimating coefficients of the transfer
15 function of the propagation channel in the respective spectral lines, means for subtracting the spectral lines of the unwanted zero Doppler effect signals deduced from the coefficients of the estimated transfer function from the spectral lines of the symbol signal, and means for
20 synthesizing the spectral lines produced by the subtraction means into the filtered signal. This first embodiment eliminates unwanted zero Doppler effect signals, i.e. essentially signals caused by direct and multiple paths from a given emitter.

25 To better characterize the transfer function of the propagation channel independently of signals backscattered from targets, the spectral lines of the unwanted signals are estimated for each symbol and averaged over each frame in the detecting means before
30 being subtracted from the spectral lines of the symbol

signal. This first embodiment provides then a means for estimating an emitted signal replica as a function of the spectral lines of the unwanted zero Doppler effect signals, the estimated replica being correlated to the
5 filtered signal in the Doppler-distance correlation means.

According to a second embodiment for a radar having several receive channels, in which the shaping means
10 comprises a plurality of receiver means and shapes a plurality of signals received by the receiver means in the form of digital symbol signals, the filtering means comprises means for producing groups of spectral lines of the symbol signals respectively corresponding to the
15 orthogonal carriers, means for estimating covariance matrices each depending on products of spectral lines two by two in a group relating to a respective carrier, means for deducing inverse matrices of the covariance matrices, means for filtering the groups of spectral lines
20 respectively relating to the carriers by multiplying the groups of lines by the respective inverse matrices to produce filtered groups of spectral lines and means for synthesizing the filtered groups of spectral lines into filtered symbol signals including essentially signals
25 backscattered by targets applied to the correlation means. This second embodiment also eliminates scatterers other than COFDM coded correlated signals.

To better characterize the transfer function of the propagation channel, spectral line products on which the
30 covariance matrices are dependent depend on spectral

lines of symbols and are averaged over each frame.

The second embodiment provides also a means for estimating an emitted signal replica as a function of the spectral lines of one of the symbols signals by
5 estimating coefficients of the transfer function of the propagation channel in the unwanted zero Doppler effect signals, the estimated replica being correlated to the filtered symbol signals in the correlation means.

10 Other features and advantages of the present invention will become more clearly apparent on reading the following description of a plurality of preferred embodiments of the invention given with reference to the corresponding accompanying drawings, in which:

15 - FIG. 1 is a timing diagram of successive symbols in a emitted COFDM signal;

- FIG. 2 is a timing diagram of a COFDM signal frame;

20 - FIG. 3 is a diagram of the following paths between an emitter and a receiver: a direct path, multiple paths, and a path reflected by a target;

- FIG. 4 shows the outputs of the various contributors (direct path, multiple paths, target, noise) to the output of the correlator;

25 - FIG. 5 is a schematic block diagram of a radar receiver with one antenna conforming to a first embodiment of the invention;

- FIG. 6 is a schematic block diagram of a radar receiver with a plurality of antennas conforming to a
30 second embodiment of the invention;

- FIG. 7 is a radiation diagram of the antenna array of the second embodiment of a radar receiver.

The main characteristics of COFDM
5 radiocommunication signals are summarized hereinafter, with reference to FIG. 1.

These baseband signals are emitted in symbol periods T'_S . A message contained in each of the emitted symbols is carried by a large number of sinusoids emitted
10 simultaneously. These sinusoids constitute sub-carriers, referred to hereinafter for simplicity as "carriers", and are phase coded or amplitude coded. The carrier frequencies f_1 to f_K are equidistant at intervals $1/T_S$. Each symbol S_i emitted is the result of summing carriers
15 as follows during the time period T'_S ($T'_S > T_S$):

$$S_i = \sum_{k=1}^{k=K} C_{k,i} e^{2j\pi kt/T_S}$$

in which j designates the square root of -1 ($j^2 = -$
20 1) and t designates time.

The carriers at frequencies $f_k = k/T_S$ with $1 \leq k \leq K$ are therefore orthogonal to the duration T_S . $\Delta = T'_S - T_S$ designates the guard time.

Over an analysis period T_S , the COFDM signals thus
25 constitute a band signal K/T_S comprising a spectrum of K lines equidistant at the frequency step $1/T_S$ and each of width $1/T_S$. The carriers are modulated individually, for example using a four-state phase code expressed by the complex coefficients $C_{k,i}$, belonging to the alphabet
30 $(1+j, 1-j, -1+j, -1-j)$.

In practise, a data message may occupy only a few frequencies f_k over a few symbol period T'_S in the time-division and frequency-division multiplex comprising K frequencies and I time slots.

5 On emission, the symbols are organized into frames. Each frame shown in FIG. 2 comprises I symbols S_1 to S_I . The first symbol S_1 of the frame is a "null" symbol carrying no information and consists of the modulation carrier frequency F_0 of the emitted signal. The symbol S_1
10 is used for frame synchronization, i.e. to provide a time reference. The second symbol S_2 of the frame is used in the radar receiver, among other things, to learn the propagation channel, and contains the K carriers or sinusoids at the frequencies f_1 to f_K having
15 predetermined phases in the emitter. These K carriers in the symbol S_1 are used in the radar receiver to estimate the emitted signal, each carrier f_k serving as a frequency and phase reference. Thanks to the first two symbols, at least the emission parameters F_0 , T_S and f_1
20 to f_K can be acquired in this way. The other symbols S_3 to S_I are intended to carry one or more data messages occupying each symbol partially or otherwise.

In radiocommunication, the received symbols are recovered with the aid of a frequency analysis of the
25 received COFDM signals received over the time period T_S . The frequencies emitted being orthogonal in the Fourier transform (FFT) sense, each of the carriers is demodulated to reconstitute the information.

In practise, the orthogonality of the frequencies
30 emitted is degraded by the following forms of

interference:

- intersymbol intracarrier interference:
overlapping of signals coded differently or analysis time
 T_S not suitable for the code;

5 - intersymbol intercarrier interference: non-
orthogonal signals over the analysis time T_S ;

- intrasymbol intracarrier interference:
overlapping of signals coded differently;

- intrasymbol intercarrier interference: non-
10 stationary signals.

These interferences are associated with the
multiple paths in the propagation channel between the
emitter and the receiver.

The addition on the guard time Δ to each symbol
15 period T'_S eliminates all kinds of interference if the
guard time Δ is greater than the temporal spreading of
the propagation channel caused by multiple paths: there
is then for each duration T'_S a range, of length T_S , in
which all the received multipath signals are coded
20 identically.

The processing of the received signals including
Doppler-distance correlation, the ambiguity function of
the COFDM signals, and in particular of their secondary
lobes, must be studied. The secondary lobes of the
25 ambiguity function associated with the COFDM waveform are
relatively uniform in the distance-Doppler plane, and
their level relative to the main lobe is $-10 \cdot \log_{10}(I.K)$.
The secondary lobes are lower at the base of the main
lobe.

30 The analysis of the conventional radar balance

shows that in general the energy contained in the secondary lobes associated with the direct path dominates over thermal noise.

Consider for example the bistatic radar shown in
5 FIG. 3 with an emitter EM radiating an electromagnetic power of $P_e G_e = 1\ 000\text{ W}$, a receiver RE antenna gain of $G_r = 10\text{ dB}$, a wavelength of $\lambda = 1\text{ m}$ corresponding to the frequency $F_0 = 300\text{ MHz}$, an receiver distance of $d = 40\text{ km}$, an emitter-target distances EM-CB and a
10 target-receiver distances CB-RE equal to 40 km , a noise factor of $F = 6\text{ dB}$, a radar equivalent surface $SER = 0\text{ dB}$, $I = 100$ symbols of duration $T_s = 1\text{ ms}$ and a guard time of $250\text{ }\mu\text{s}$ for a frame duration of 125 ms , and $K = 1\ 500$ carriers for a bandwidth of $B = 1.5\text{ MHz}$.

15 FIG. 4 shows the link balance between the emitter EM and the receiver RE.

The energy contained in the secondary lobes attached to the direct path and to the clutter caused by the multiple paths dominates over thermal noise. The
20 direct path lobe energy level is $186 - 135 = 51\text{ dB}$ above the presumed level of the target, which has a signal to thermal noise ratio of $198 - 186 = 12\text{ dB}$.

The invention therefore aims to reject effectively the direct path and clutter in a received broadband
25 signal before Doppler-distance correlation in order to detect mobile targets.

FIG. 5 shows a passive radar receiver REa in accordance with the invention for OFDM signals that
30 comprises a received signal shaping circuit 1, a

correlated signal filtering circuit 2, a correlated signal detector 3, a target determination circuit 4, and an emitted signal replica estimator 5.

5 The received signal shaping circuit 1 conventionally comprises at its input an antenna 11 and an OFDM radio frequency receiver stage 12 analogous to those for receiving OFDM radio and television broadcast signals. Following frequency conversion, the receiver
10 stage 12 digitizes the next baseband radio signal $X(t)$ received and applies it to a emission parameter and channel estimator circuit 13: super-position

$$X(t) = TD(t) + SC(t) + B(t),$$

15

where for each symbol emitted $\sum_{k=1}^{k=K} C_k e^{2j\pi kt/T_s}$,

ignoring the index i of the symbol S_i :

$$20 \quad - TD(t) = \sum_{k=1}^{k=K} H_k C_k e^{j2\pi kt/T_s}$$

designates an OFDM signal received from at least one emitter EM or possibly from a plurality of OFDM emitters on direct paths and multiple paths caused by fixed
25 reflectors RF and corresponding to clutter, as shown in FIG. 3. H_k designates a complex coefficient corresponding to the transfer function of the propagation channel EM-REa relating to these paths for the line f_k ; the signal $TD(t)$ is constituted by overlapping unwanted zero Doppler

effect signals caused by direct and multiple paths; these unwanted signals are referred to as "correlated signals";

$$- SC(t) = \sum_{k=1}^{k=K} h_c C_k e^{j2\pi k(t-\tau)/T_s} e^{-j2\pi vt}$$

5

designates a received OFDM signal caused by backscattering of emitted OFDM signals by at least one mobile target CB and thus subject to a non-zero Doppler effect; it constitutes the wanted signal to be extracted, whose power is very low compared to that of the correlated signal $TD(t)$; h_c is the narrowband transfer function of the target CB, τ denotes the path difference between the direct path and the path reflected by the target, and v is the Doppler frequency of the target;

15 - $B(t)$ designates signals received in the wanted bandwidth, called scatterers, other than the correlated OFDM signals, such as interference and thermal noise signals.

The circuit 13 estimates the parameters of the emitted signal, such as the carrier frequency F_0 and the symbol period T_s , as a function of an analysis of the first two symbols S_1 and S_2 of a frame (FIG. 2) in order to constitute a time reference. Knowing the period T'_s of the emitted signal and the duration T_s of each symbol, the temporal length of the channel is deduced from the process of synchronizing the preceding time reference, by analyzing the signal received in each guard time Δ , which is greater than the temporal channel length.

The received signal is then periodically truncated in a truncator circuit 14. The stationary portion of the

received symbols of duration T_S is recovered by subtracting the signal portion received in the guard time Δ of each period T'_S and in particular by subtracting the channel length deduced in each period.

5

Each portion of the received digital baseband signal of duration T_S is then applied to a Fourier analyzer 21 at the input of the filtering circuit 2. The analyzer produces the real and imaginary components of the received signals for each duration T_S using the Hilbert transform and analyzes them using the fast Fourier transform (FFT) to supply the frequency spectrum of each symbol S_i delivered by the truncator circuit 14. The K spectral lines SP_1 to SP_K of the symbol relating to the frequencies f_1 to f_K are applied to the detector 3 and to first inputs of a subtractor 22. The information carried by each spectral line SP_k relating to a frequency emitted independently of the other frequencies is related on the one hand to the coding of the corresponding spectral line and on the other hand to the transfer function H_k of the propagation channel.

During a recurrent learning phase in each symbol frame, the detector 3 subtracts from the spectral lines of each symbol of the frame the spectral lines of a reference signal $SR(t)$, such as:

$$SR(t) = \sum_{k=1}^{K} \bar{H}_k C_k e^{j2\pi kt/T_S} + b(t).$$

$\bar{H}_k$ is the mean of the $I-1$ coefficients of the propagation channel transfer function for the line SP_k

during the symbols S_2 to S_I of the frame, in other words virtually $|\bar{H}_K| = |H_K|$, which improves the estimate of the transfer function of the channel by making it less dependent on the instantaneous amplitude and phase variations. This averaging means that the backscattered target signal $SC(t)$ can be ignored and thus the propagation channel and the correlated signals can be characterized. The noise signal $b(t)$ designates a mean of scatterers received during the frame, consisting essentially of thermal noise with a variance lower than that of the scatterers $B(t)$.

After storing the transfer function coefficients $\bar{H}_1$ to $\bar{H}_K$ averaged over a frame, the detector 3 applies the spectral lines of the zero Doppler effect correlated signals averaged over a frame, i.e. the averaged lines of the signal $SR(t)$ depending on the coefficients $\bar{H}_1$ to $\bar{H}_K$ at K second inputs of the subtractor 22. The subtractor subtracts the lines of the signal $SR(t)$ from the spectral lines SP_1 to SP_K of the received signal relating to each symbol S_2 to S_I of the frame. The subtractor 22 then produces spectral lines of a filtered signal:

$$X'(t) = SC(t) + B(t) - b(t)$$

in which the contribution of correlated signals with zero Doppler effect caused by the direct path and the multiple paths is eliminated by subtracting $H_K - \bar{H}_K = 0$ for each respective frequency f_k .

In the filtering circuit 2, the lines of the signal $X'(t)$ are synthesized using an inverse fast Fourier transform FFT^{-1} in a synthesizer 23 which reconstitutes the stream of symbols of the digital signal $X'(t)$

comprising primarily the target backscatter signal $SC(t)$. The signal $X'(t)$ is applied to the target discriminator circuit 4.

In parallel with this, the emitted signal replica estimator 5 receives the coefficients $\bar{H}_k = H_k$ of the transfer function of the channel estimated by the detector 3 and the spectral densities of the reference signal $SR(t)$, in order to estimate a replica $Re(t)$ of the signal emitted:

10

$$Re(t) = \sum_{k=1}^{k=K} C_k e^{j2\pi kt/T_S}.$$

In a similar manner to a Doppler channel of the radar receiver disclosed in the French patent application FR 2776438, the target discriminator circuit 4 includes a Doppler-distance correlator 41. Doppler channels in the correlator 41 are assigned a predetermined frequency offset relative to each other because of the Doppler effect. The Doppler channels constitute a plurality of speed cases of the filtered signal $X'(t)$ by changes of frequency and are each correlated, for each carrier f_1 to f_K , to the emitted signal replica $Re(t)$. After summation, an angular focusing circuit 42 determines mobile targets angular position cases. Finally, a constant false alarm processing (CFAP) circuit 43 extracts "plots" relating to position and speed data on searched-for mobile targets.

Although this first embodiment has been described for a bistatic radar, it can be applied to a monostatic radar. It can also be used in a radar with a plurality of receive antennas and therefore a plurality of filtering

operations by subtracting spectral lines depending on transfer function coefficients $\bar{H}_1$ to $\bar{H}_K$ for the K spectral lines relating to each receive antenna in the filtering circuit 2, which produces as many filtered
5 signals $X'(t)$ as there are antennas to be processed in parallel in the Doppler channels of the circuit 4, as in the circuit 4b shown in FIG. 6.

In the embodiment described above with reference to
10 FIG. 5, the received signal $X(t)$ is processed to eliminate the contribution of the direct path and the multiple paths between at least one COFDM emitter EM and the receiver REa. However, scatterers, such as interference and thermal noise in the wanted frequency
15 band, are not eliminated in the signal $X'(t)$ processed by the correlator 41.

In a second embodiment, shown in FIG. 6, the radar receiver REb aims to eliminate all correlated unwanted signals and scatterers.

20 The receiver REb for COFDM signals comprises a plurality of receiver antennas 11_1 to 11_N connected to a plurality of receivers 12_1 to 12_N , respectively, with $N \geq 2$. The radar receiver REb has a structure identical to that of the receiver REa shown in FIG. 5, but with N
25 parallel receive channels between the circuits 13b, 14b, 2b and 4b respectively associated with the antennas 11_1 to 11_N .

The detector 3 and the replica estimator 5 are not modified. The detector 3 produces the reference signal
30 $SR(t)$ relating to one of the antennas 11_1 to 11_N , for

example as a function of the K spectral lines SP_{11} to SP_{K1} of the first channel connected to the antenna 11_1 , which are delivered by the spectrum analyzer 21b which analyzes N received symbol streams supplied by the truncator circuit 14b. The estimator 5 produces a narrowband emitted signal replica $Re(t)$ applied to N Doppler-distance correlators 41b in the target discriminator circuit 4.

The receiver REb essentially differs from the receiver REa in the narrowband filtering circuit 2b which eliminates unwanted zero Doppler effect signals $TD(t)$ and scatterers $B(t)$.

The filtering circuit 2b comprises, between the spectrum analyzer 21b producing in parallel K spectral lines for each symbol of the received signals $X_1(t)$ to $X_N(t)$ and the Fourier synthesizer 23b supplying N filtered signals $X'_1(t)$ to $X'_N(t)$ to the correlators 41b, successively a covariance matrix calculation module 24b, an inverse covariance matrix calculation module 25b, and a filtering module 26b.

For each line SP_k at the frequency f_k delivered by the analyzer 21b, the module 24b estimates an $N \times N$ covariance matrix R_k in which a row of given rank n is made up of products of the spectral line SP_{kn} received relating to a given antenna 11_n by the conjugates of the received spectral lines SP_{k1} to SP_{kN} relating to the N antennas 11_1 to 11_N , the products being averaged for symbols having a predetermined duration, with $1 \leq n \leq N$. The predetermined duration for averaging said products is preferably significantly longer than the symbol period

T_S , and is typically the duration of a frame, i.e. products averaged over $I-1$ symbols S_2 to S_I . Averaging said products over a large number of symbols, i.e. over around a hundred symbols constituting the frame, decorrelates the target signal from the unwanted signals received by the antennas at a high level. The mutual orthogonal relationship of the spectral lines makes the covariance matrices R_1 to R_K independent of the coding.

The module 25b then deduces the inverse matrices R_1^{-1} to R_K^{-1} of the K covariance matrices and stores them. These K $N \times N$ inverse matrices serve as K filters in the module 26b for filtering K respective groups each of N spectral lines $SP_{11} - SP_{1N}$ to $SP_{K1} - SP_{KN}$ delivered by the analyzer 21b. Each group of N spectral lines SP_{k1} to SP_{kN} received for a given frequency f_k is thus filtered by a filter which for each symbol supplies the product of the column vector composed of N received spectral lines $SP_{k1} - SP_{kN}$ for that symbol and the frequency f_k by the inverse matrix R_k^{-1} . The N groups each of K filtering signals supplied by the filtering module 26b are then applied to the synthesizer 23b which delivers N symbol time signals $X'_1(t)$ to $X'_N(t)$ in the correlators 41b. In the target discriminator circuit 4b, the N correlators 41b are followed by angular focusing circuits 42b and a CFAP circuit 43b.

The radiation diagrams of the signals $X'_1(t)$ to $X'_N(t)$ has blind axes, i.e. "holes" in the directions for receiving correlated OFDM signals and scatterers, as shown in FIG. 7.

CLAIMS

1. A radar receiver processing a radio signal received via a propagation channel (EM-RE) and comprising
5 frames of symbols each emitted on coded orthogonal carriers ($f_1 - f_K$), comprising shaping means (1; 1b) for converting the received signal into a digital symbol signal (X, S_1-S_I) and Doppler-distance correlation means (4; 4b) for discriminating mobile targets, characterized
10 in that it comprises filtering means (2; 2b) for eliminating in the symbol signal (X) at least unwanted zero Doppler effect signals (TD) in order to apply a filtered signal (X') including essentially signals (SC) backscattered by targets (CB) to the correlation means
15 (4; 4b).

2. A radar receiver according to claim 1, wherein the filtering means (2) comprises means (21) for producing spectral lines ($SP_1 - SP_K$) of the symbol signal
20 (X) corresponding to the orthogonal carriers ($f_1 - f_K$), means (3) for detecting the unwanted zero Doppler effect signals (TD) by estimating coefficients of the transfer function ($\bar{H}_1 - \bar{H}_K$) of the propagation channel respectively in the spectral lines, means (22) for
25 subtracting the spectral lines of the unwanted zero Doppler effect signals (TD) deduced from the coefficients of the estimated transfer function from the spectral lines of the symbol signal ($SP_1 - SP_K$), and means (23) for synthesizing the spectral lines produced by the
30 subtraction means (22) into the filtered signal (X').

3. A radar receiver according to claim 2, wherein the spectral lines of the unwanted signals (TD) are estimated for each symbol and averaged over each frame in the detecting means (3) before being subtracted from the spectral lines ($SP_1 - SP_K$) of the symbol signal.

4. A radar receiver according to claim 2 or 3, comprising means (5) for estimating an emitted signal replica (Re) as a function of the spectral lines of the unwanted zero Doppler effect signals (TD), the estimated replica being correlated to the filtered signal (X') in the Doppler-distance correlation means (4).

5. A radar receiver according to claim 1, wherein the shaping means (1b) comprises a plurality of receiver means ($11_1, 12_1$ to $11_N, 12_N$) and shapes a plurality of signals received by the receiver means in the form of digital symbol signals ($X_1 - X_N$), characterized in that the filtering means (2b) comprises means (21b) for producing groups of spectral lines ($SP_{11} - SP_{1N}$ to $SP_{K1} - SP_{KN}$) of the symbol signals ($X_1 - X_N$) respectively corresponding to the orthogonal carriers ($f_1 - f_K$), means (24b) for estimating covariance matrices ($R_1 - R_K$) each depending on products of spectral lines (SP_{k1} to SP_{kN}) two by two in a group relating to a respective carrier (f_k), means (25b) for deducing inverse matrices (R_1^{-1} to R_K^{-1}) of the covariance matrices, means (26b) for filtering the groups of spectral lines respectively relating to the carriers ($f_1 - f_K$) by multiplying the

groups of lines by the respective inverse matrices to produce filtered groups of spectral lines and means (23b) for synthesizing the filtered groups of spectral lines into filtered symbol signals $(X'_1 - X'_N)$ including
5 essentially signals backscattered by targets (CB) applied to the correlation means (4b).

6. A radar receiver according to claim 5, wherein the spectral line products on which the covariance
10 matrices $(R_1 - R_K)$ are dependent depend on spectral lines of symbols and are averaged over each frame.

7. A radar receiver according to claim 5 or 6, comprising means (3, 5) for estimating an emitted signal
15 replica (Re) as a function of the spectral lines $(SP_{11} - SP_{K1})$ of one of the symbols signals (X_1) by estimating coefficients of the transfer function of the propagation channel in the unwanted zero Doppler effect signals, the estimated replica being correlated to the
20 filtered symbol signals $(X'_1 - X'_N)$ in the correlation means (4b).