Underwater Acoustic Model-Based Signal Processing

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Abstract—An underwater acoustic model-based signal processing algorithm is presented. Its performance was evaluated via computer simulation of various test cases and compared to that predicted by theory. The model-based algorithm, which is used in conjunction with an FFT beamformer for planar arrays, computes phase weights that correct for deterministic, ocean medium, phase effects due to ray bending as a signal propagates in the inhomogeneous ocean medium whose index of refraction (sound-speed profile) is a function of depth.

The performance of the model-based signal processing algorithm was evaluated in the context of an underwater acoustic communication problem in order to determine the impact of the model-based algorithm on the probability of detecting single rectangular-envelope continuous wave (CW) and linear frequency modulated (LFM) pulses as a function of the input signal-to-noise power ratio at a single element in the receive array for a given probability of false alarm. Preliminary results for various test cases show significant increases in performance for a correlator receiver.

List of Often-Used Parameters

- $c'_i, d'_i$: complex weights in the $X$ and $Y$ directions, respectively, associated with the transmit array
- $c_m, d_m$: complex weights in the $X$ and $Y$ directions, respectively, associated with the receive array.
- $c(y)$: speed of sound (in meters per second) as a function of depth $y$
- $d'_X, d'_Y$: interelement spacings in meters in the $X$ and $Y$ directions, respectively, associated with the transmit array
- $d_X, d_Y$: interelement spacings in meters in the $X$ and $Y$ directions, respectively, associated with the receive array.
- $DMEDIA$: logical variable; if $DMEDIA = \text{TRUE}$, then model-based phase weights are used to correct for deterministic, inhomogeneous ocean medium, phase effects
- $f_c$: carrier frequency in hertz
- $g$: constant gradient (in seconds$^{-1}$) of linear sound-speed profile
- $H(f, m, n)$: overall system complex frequency response at element $(m, n)$ in the receive array
- $H_M(f, f_T; y)$: random, time-invariant space-variant, ocean medium transfer function
- $k(y)$: wave number (in radians per meter) as a function of depth $y$
- $M', N'$: total odd number of elements in the $X$ and $Y$ directions, respectively, associated with the transmit array
- $M, N$: total odd number of elements in the $X$ and $Y$ directions, respectively, associated with the receive array
- $n_D(y), n_{NR}(y)$: deterministic and normalized random components of the index of refraction, respectively, as a function of depth $y$
- $\theta_{MD}(f, f_T, y)$, $\theta_{MR}(f, f_T, y)$: deterministic and random components of the ocean medium phase function
- $\theta_m(f), \phi_n(f)$: phase weights in the $X$ and $Y$ directions, respectively, associated with element $(m, n)$ in the receive array
- $\phi_{MD}(f, n)$: model-based phase weight in the $Y$ direction associated with the receive array
- $\sigma(y)$: standard deviation of the random component of the index of refraction as a function of depth $y$
- $\text{STEER}$: logical variable; if $\text{STEER} = \text{TRUE}$, then standard phase weights based on line of sight geometrical considerations alone are used
- $u_o = \sin \theta_o \cos \psi_o$, $v_o = \sin \theta_o \sin \psi_o$: direction cosines with respect to the $X$ and $Y$ axes, respectively, representing initial directions of wave propagation at the transmit array (see Fig. 1)
- $X_T, Y_T, Z_T$: rectangular coordinates of the center of the transmit array
- $X_R, Y_R, Z_R$: rectangular coordinates of the center of the receive array
- $y_0$: depth (in meters) of point sound source

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I. Introduction

Model-based signal processing is described by Mendel [1] as exploiting "the detailed physics of a problem area to construct precise and tractable mathematical formulations of appropriate signal processing algorithms. Particularly in geophysical signal processing, a close coupling between the physics and the signal processing is essential for real progress." Other researchers, for example, [2]–[10], although not using the term "model-based signal processing" explicitly, have been proponents of treating the ocean medium as an underwater acoustic communication channel in addition to advocating the philosophy of a close coupling between physics and signal processing.

The purpose of this paper is to present an underwater acoustic model-based signal processing algorithm and to evaluate its performance versus that predicted by theory via computer simulation of various test cases. The model-based signal processing algorithm is used in conjunction with a three-dimensional FFT beamformer for planar arrays [11]. The model-based algorithm computes phase weights that correct for deterministic, ocean medium, phase effects due to ray bending as a signal propagates in the inhomogeneous ocean medium whose index of refraction (sound-speed profile) is a function of depth. The point to be made is that in order to detect a signal propagating in an inhomogeneous medium, traditional beamsteering is not sufficient to co-phase all of the output electrical signals from each element in an array. It will be shown that additional beamsteering must be done using the model-based signal processing algorithm to ensure that all of the output electrical signals from each element in an array are co-phased and, hence, that the theoretical value of array gain possible is in fact achieved.

The performance of the model-based signal processing algorithm was evaluated in the context of an underwater acoustic communication problem (see Fig. 1). In order to drive the algorithm, computer simulated output electrical signals, based on derived mathematical models, were generated at each element in a receive planar array of point sources. The output electrical signals depend on the frequency spectrum of the transmitted electrical signal, the far-field beam patterns of the transmit and receive planar arrays, and the time-invariant space-variant random transfer function of the ocean volume which was derived using the Wentzel, Kramers, and Brillouin (WKB) approximation [12]. The transfer function was time invariant because motion was not considered. The ocean volume was characterized by a one-dimensional random index of refraction (sound-speed profile) which was a function of depth. The index of refraction was decomposed into a deterministic component and a zero mean random component. Both single rectangular-envelope CW and LFM pulses were used as transmitted electrical signals. The mathematical model of the output electrical signal at each element in a receive planar array of point sources is discussed in Section II of this paper.

The computer simulated output electrical signals from each element in the receive planar array were first processed by a three-dimensional FFT beamformer [11] that was capable of utilizing either standard phase weights alone or the sum of standard and model-based phase weights. The composite output signal from the FFT beamformer was then processed by a correlator receiver, followed by a magnitude square operation, and finally by a Neyman-Pearson test (see Fig. 3) in order to determine the impact of the model-based algorithm on the probability of detecting various transmitted electrical signals as a function of the input signal-to-noise power ratio at a single element in the receive array for a given probability of false alarm. Preliminary results obtained by processing the computer simulated output signals with a three-dimensional FFT beamformer alone, without the model-based signal processing algorithm and correlator receiver, were reported by Ziomek and Vos [13]. The model-based signal processing algorithm is discussed in Section III of this paper. The various test cases and computer simulation results are discussed in Section IV, and Section V is devoted to a discussion of conclusions.

II. Output Electrical Signal

Assume that the transmit aperture depicted in Fig. 1 is a planar array of \( M' \times N' \) (odd) complex-weighted point sources, centered at \( (x_\circ, y_\circ, z_\circ) = (x_R, y_R, z_R) \) and parallel to the \( XY \) plane. Similarly, assume that the receive aperture depicted in Fig. 1 is a planar array of \( M \times N \) (odd) complex-weighted point sources, centered at \( (x = x_R, y = y_R, z = z_R) \) and parallel to the \( XY \) plane. Therefore, it can be shown that the random output electrical signal \( y(t, x, y, z) \) at each element \( (m, n) \) in the receive planar array is given by [14]

\[
y(t, x, y, z) = \sum_{m=-(M-1)/2}^{(M-1)/2} \sum_{n=-(N-1)/2}^{(N-1)/2} c_m d_n y(t, m, n) \cdot \delta[x - (x_R + m d_x)]
\]

\[
y(t, x, y, z) = \sum_{m=-(M-1)/2}^{(M-1)/2} \sum_{n=-(N-1)/2}^{(N-1)/2} c_m d_n y(t, m, n) \cdot \delta[x - (x_R + m d_x)]
\]

\[
f(t, x, y, z) = \sum_{m=-(M-1)/2}^{(M-1)/2} \sum_{n=-(N-1)/2}^{(N-1)/2} c_m d_n y(t, m, n) \cdot \delta[x - (x_R + m d_x)]
\]
where
\[ y(t, m, n) = F_f^{-1} \{ Y(f, m, n) \} \]
\[ = \int_{-\infty}^{\infty} Y(f, m, n) \exp(j2\pi ft) \, df \quad (2-2) \]
is the output electrical signal from element \((m, n)\) in the receive planar array before the application of the complex weights \(c_m\) and \(d_n\),
\[ Y(f, m, n) = X(f) H(f, m, n) \quad (2-3) \]
\[ X(f) \] is the complex frequency spectrum of the transmitted electrical signal,
\[ H(f, m, n) = f^2 \sum_{\nu_0=a_0}^{+1} \sum_{q=-N^{-1}/2}^{N^{-1}/2} c^{-2}(q) d_q^* H_M(f, [f\nu_0/c(q)]; \nu_R + nd_f) \]
\[ \cdot \exp\left[ -j\theta_M(f, f_Y; \nu) \right] S_x(f, u_o, q) \]
\[ \cdot \exp\left[ +j\phi(f, u_o, \nu_0, q, m) \right] du_o \, dv_o \quad (2-4) \]
can be thought of as the overall system complex frequency response,
\[ H_M(f, f_Y; \nu) = A_M(f, f_Y; \nu) \exp\left[ +j\theta_M(f, f_Y; \nu) \right] \quad (2-5) \]
is the random, time-invariant space-variant transfer function of the ocean medium based on the WKB approximation where \([12]\)
\[ A_M(f, f_Y; \nu) = (2\pi f_Y)^{-1/2} \quad (2-6) \]
\[ \theta_M(f, f_Y; \nu) = \theta_{MD}(f, f_Y; \nu) + \theta_{MR}(f, f_Y; \nu) \quad (2-7) \]
\[ \theta_{MD}(f, f_Y; \nu) = -\left[ k_0^2/(4\pi f_Y) \right] \int_{\nu_0}^{\nu} \left[ n_D^2(\xi) - 1 \right] d\xi \quad (2-8) \]
is the deterministic or average component of the phase function,
\[ \theta_{MR}(f, f_Y; \nu) = -\left[ k_0^2/(2\pi f_Y) \right] \]
\[ \cdot \int_{\nu_0}^{\nu} n_D(\xi) \sigma(\xi) n_{NR}(\xi) d\xi \quad (2-9) \]
is the random component, \(n_D(\nu)\) is the deterministic or average component of the index of refraction, \(n_{NR}(\nu)\) is the zero mean, unit variance, normalized random component, and \(\sigma(\nu)\) is the standard deviation of the zero mean random component of the index of refraction \(n_R(\nu)\)
\[ = \sigma(\nu) n_{NR}(\nu) \]
\[ S_X(f, u_o, q) = \sum_{i=-(M-1)/2}^{(M-1)/2} c_i^* \exp\left\{ +j2\pi \left[ f u_o/c(q) \right] i d_f \right\} \quad (2-10) \]
is the far-field beam pattern of the transmit array in the \(X\) direction at a source depth of
\[ y_o = y_T + q d_f \quad (2-11) \]
\[ \phi(f, u_o, \nu_0, q, m) = -k(q) \left\{ u_o \Delta X_m + \left[ 1 - (u_o^2 + v_o^2) \right]^{1/2} \Delta Z \right\} \quad (2-12) \]
is the speed of sound in meters per second at a source depth of
\[ y_o = y_T + q d_f \quad (2-13) \]
is the corresponding wave number in radians per meter,
\[ \Delta X_m = x_R - x_T + m d_x \quad (2-15) \]
\[ \Delta Y_m = y_R - y_T + m d_y \quad (2-16) \]
\[ \Delta Z = z_R - z_T \quad (2-17) \]
is the transmitted spatial frequency in cycles per meter in the \(Y\) direction,
\[ u_o = \sin \theta_o \cos \psi_o \quad (2-19) \]
is the direction cosine with respect to the \(X\) axis,
\[ v_o = \sin \theta_o \sin \psi_o = \cos \beta_o \quad (2-20) \]
is the direction cosine with respect to the \(Y\) axis,
\[ a_o = \left[ 1 - n_{NR}(\nu_R + nd_f) \right]^{1/2} \quad (2-21) \]
is the lower limit of integration with respect to direction cosine \(\nu_0, c_i^*\) and \(d_q^*\) are the complex weights in the \(X\) and \(Y\) directions, respectively, associated with the transmit array; \(d'_X\) and \(d'_Y\) are the interelement spacings in meters in the \(X\) and \(Y\) directions, respectively, associated with the transmit array; \(c_m\) and \(d_n\) are the complex weights in the \(X\) and \(Y\) directions, respectively, associated with the receive array; \(d_k\) and \(d_d\) are the interelement spacings in meters in the \(X\) and \(Y\) directions, respectively, associated with the receive array. The direction cosines \(u_o\) and \(v_o\) and, therefore, the angles \(\theta_o, \psi_o, \beta_o\) represent initial directions of wave propagation at the transmit array (see Fig. 1).

The overall system complex frequency response \(H(f, m, n)\) given by (2-4) can be simplified by evaluating the integral with respect to direction cosine \(u_o\) using the method of stationary phase \([15]-[17]\). If \(v_o < 1\) \((\beta_o > 0^\circ)\) and \(\Delta Z \gg \Delta X_m\), then the overall system complex
The frequency response can be expressed as

\[ H(f, m, n) = f^2 \sum_{k=n}^{\infty} c^{-1/2}(q) \cosh^{-1}(v_0^2) \exp \left\{ -\frac{1}{2} (f + m) \right\} \delta_{2/3} \left( \frac{c_0^2}{(4\pi f_s)} \right) \left\{ \frac{(c_0 / g) [n_R(y) - 1]}{y - y_0} \right\} \]  

(2-31)

The random phase function given by (2-9) can also be simplified if the random component of the index of refraction is not a function of depth. If \( n_R(y) = \sigma(y) n_{NR}(y) = n_R \),

where \( n_R \) is a zero mean random variable with variance \( \sigma^2 \), then (2-9) reduces to

\[ \theta_{MR}(f, f_s; y) = -\frac{k_0^2}{(2\pi f_s)} \int_{y_0}^{y} n_R(\xi) d\xi. \]  

(2-33)

Since

\[ \int_{a}^{b} \frac{dy}{a + by} = \frac{1}{b} \ln (a + by), \]  

(2-34)

substituting (2-27) into (2-33) yields the following closed-form expression for the random phase function of the ocean medium:

\[ \theta_{MR}(f, f_s; y) = \frac{k_0^2}{(2\pi f_s)} (c_0 / g) n_R \ln [n_D(y)] \]  

(2-35)

or, since \( n_R = \sigma n_{NR} \) where \( n_{NR} \) is a zero mean, unit variance random variable,

\[ \theta_{MR}(f, f_s; y) = \frac{k_0^2}{(2\pi f_s)} (c_0 / g) \sigma n_{NR} \ln [n_D(y)]. \]  

(2-36)

The magnitude of the constant standard deviation \( \sigma \) of the random component of the index of refraction \( n_R \) is on the order of 10^{-4} [18], [19].

One further step of simplification is possible. Since the transmitted electrical signal \( x(t) \) is, in general, an amplitude and angle modulated carrier, it can be represented as

\[ x(t) = \Re \{ \hat{x}(t) \exp (+j2\pi f_s t) \} \]  

(2-37)

where

\[ \hat{x}(t) = a(t) \exp [ +j\theta(t) ] \]  

(2-38)

is the baseband complex envelope of the real bandpass signal \( x(t) \), \( a(t) \) and \( \theta(t) \) are real amplitude and angle modulating signals, respectively, and \( f_s \) is the carrier frequency in hertz [20]. The relationship between \( X(f) \) and \( X(\hat{x}(f)) \) is given by [20]

\[ X(f) = 0.5 \left\{ \hat{X}(f - f_c) + \hat{X}^*[-(f + f_c)] \right\} \]  

(2-39)

where

\[ X(f) = F_i\{ x(t) \} \]  

(2-40)

and

\[ \hat{X}(f) = F_i\{ \hat{x}(t) \} \]  

(2-41)
If we represent the baseband complex envelope \( \hat{x}(t) \) by a finite Fourier series during the time interval \( |t| \leq T_o/2 \), that is, if

\[
\hat{x}(t) = \sum_{k=-K}^{K} c_k \exp (+j2\pi kf_o t), \quad |t| \leq T_o/2, \tag{2-42}
\]

then

\[
\hat{X}(f) = \sum_{k=-K}^{K} c_k \delta(f - kf_o) \tag{2-43}
\]

where

\[
c_k = \frac{1}{T_0} \int_{-T_o/2}^{T_o/2} \hat{x}(t) \exp (-j2\pi kf_o t) \, dt \tag{2-44}
\]

is the complex Fourier series coefficient for the \( k \)th harmonic, \( f_o = 1/T_o \) is the fundamental frequency in hertz, \( T_o \) is the fundamental period in seconds, and \( K \) is the highest harmonic used in the finite Fourier series representation. Substituting (2-3), (2-39), and (2-43) into (2-2) yields

\[
y(t, m, n) = 0.5 \sum_{k=-K}^{K} c_k H(f_c + kf_o, m, n) \quad \exp [ +j2\pi (f_c + kf_o)t ]
\]

\[
+ 0.5 \sum_{k=-K}^{K} c_k^* H[-(f_c + kf_o), m, n] \quad \exp [ -j2\pi (f_c + kf_o)t ].
\tag{2-45}
\]

Since it can be shown that

\[
H[-(f_c + kf_o), m, n] = H^*(f_c + kf_o, m, n) \tag{2-46}
\]

because the complex weights used in the transmit planar array obey the following symmetry:

\[
c_{i-q} = (c_i)^*, \tag{2-47}
\]

and

\[
d_{i-q} = (d_q)^*, \tag{2-48}
\]

that is, the amplitude weights are an even function of the indexes \( i \) and \( q \), and the phase weights are an odd function of \( i \) and \( q \); (2-45) reduces to

\[
y(t, m, n) = \text{Re} \left\{ \hat{y}(t, m, n) \exp (+j2\pi f_c t) \right\} \tag{2-49}
\]

where

\[
\hat{y}(t, m, n) = \sum_{k=-K}^{K} c_i H(f_c + kf_o, m, n)
\]

\[
\quad \exp (+j2\pi kf_o t) \tag{2-50}
\]

is the baseband complex envelope of the output electrical signal from element \((m, n)\) in the receive planar array before the application of the complex weights \( c_m \) and \( d_n \).

The computer simulated output electrical signals used to drive the model-based signal processing algorithm were obtained by evaluating

\[
\hat{r}(t, m, n) = \hat{y}(t, m, n) + \hat{n}(t, m, n),
\]

\[
m = -(M - 1)/2, \cdots, 0, \cdots, (M - 1)/2
\]

\[
n = -(N - 1)/2, \cdots, 0, \cdots, (N - 1)/2 \tag{2-51}
\]

where \( \hat{r}(t, m, n) \) is the baseband complex envelope of the received signal at element \((m, n)\), \( \hat{y}(t, m, n) \) is given by (2-50) where \( H(f, m, n) \) is given by (2-22), \( H_M(f, f_y; y) \) is given by (2-5) through (2-7), (2-31), and (2-36); and \( \hat{n}(t, m, n) \) is the baseband complex envelope of the zero mean, Gaussian, random noise. Note that the results presented in this paper are based on setting \( \sigma = 0 \) and, as a result, \( \theta_{MD}(f, f_y; y) \), as given by (2-36), equal to zero.

III. A MODEL-BASED SIGNAL PROCESSING ALGORITHM

The model-based signal processing algorithm about to be presented is meant to be used in conjunction with a three-dimensional FFT beamformer for planar arrays. Each of the computer simulated output electrical signals \( \hat{r}(t, m, n) \) given by (2-51) were processed by the scheme illustrated in Fig. 2, where QD refers to a quadrature demodulator. The quadrature demodulator (QD) is shown for completeness since, in practical signal processing applications, a QD is used to obtain the baseband complex envelope \( \hat{r}(t, m, n) \) from the real bandpass signal \( r(t, m, n) \) [21], [22].

The complex weights \( c_m \) and \( d_n \) can be expressed as

\[
c_m = a_m \exp (+j\theta_m) \tag{3-1}
\]

and

\[
d_n = b_n \exp (+j\phi_n) \tag{3-2}
\]

where \( a_m \) and \( b_n \) are real amplitude weights and \( \theta_m \) and \( \phi_n \) are real phase weights. Substituting (2-11) and (2-18) into (2-31) and evaluating (2-31) at \( y = y_R + nd_y \) yields

\[
\theta_{MD}(f, [fT_o/c(q)]; y_R + nd_y)
\]

\[
= \left[ k_o/(2\nu_o) \right] \left[ (c_o/g)[n_R(y_R + nd_y) - 1] + \Delta Y_{eq} \right] \tag{3-3}
\]

which represents the deterministic angle modulation performed by the ocean medium on the transmitted electrical signal as a function of depth. Since we have a mathematical model of the effect the medium has on the phase of the transmitted electrical signal, we should be able to compensate for the medium via proper signal processing at the receive array. Therefore, using the form of (3-3), the model-based signal processing algorithm for the phase weights \( \theta_m \) and \( \phi_n \) is given by the following set of equations:

\[
\theta_m(f) = -2\pi f_k md_k, \quad m = -(M - 1)/2, \cdots, 0, \cdots, (M - 1)/2 \tag{3-4}
\]
are the phase weights in the $X$ direction, and
\[ \phi_i(f) = -2\pi f_n y_i + \phi_{MD}(f, n), \]
\[ n = -(N - 1)/2, \cdots, 0, \]
\[ \cdots, (N - 1)/2 \]  \hspace{1cm} (3-5)
are the phase weights in the $Y$ direction where
\[ f'_x = -u_B f / c(y_T), \]  \hspace{1cm} (3-6)
\[ f'_y = -v_B f / c(y_R), \]  \hspace{1cm} (3-7)
\[ \phi_{MD}(f, n) = -\left[k(y_T)/(2v_B)\right]\left[\left[c(y_T)/g\right] \cdot \left[n_y(y_R + nd_y) - 1 + \Delta Y_n\right]\right], \]  \hspace{1cm} (3-8)
\[ k(y_T) = 2\pi f / c(y_T), \]  \hspace{1cm} (3-9)
\[ f = f_c + kf_o, \quad k = -K, \cdots, 0, \cdots, K \]  \hspace{1cm} (3-10)
\[ n_y(y_R + nd_y) = c(y_T)/[c(y_T) + g\Delta Y_n], \]  \hspace{1cm} (3-11)
\[ \Delta Y_n = y_R - y_T + nd_y, \]  \hspace{1cm} (3-12)
\[ u_B \] and $v_B$ are the direction cosines in the $X$ and $Y$ directions, respectively, associated with the direction in which the transmit beam pattern is steered, $c(y_T)$ and $c(y_R)$ are the speeds of sound in meters per second at the centers of the transmit and receive arrays, respectively, $g$ is the constant gradient in seconds$^{-1}$ of the linear sound-speed profile, $f_c$ is the carrier frequency in hertz of the transmitted amplitude and angle modulated carrier, $f_o$ is the fundamental frequency in hertz of the finite Fourier series representation of the complex envelope of the transmitted electrical signal, and $K$ is the highest harmonic used in the finite Fourier series. In our problem, the transmit beam pattern was always steered toward the center of the receive array.

Equation (3-4) and the first term in (3-5) are standard phase weights for planar arrays based on line of sight geometrical considerations alone [23]. However, the second term in (3-5), which is given by (3-8), is a model-based phase weight. It compensates for deterministic medium phase effects due to ray bending. Note that the spatial frequency $f'_x$ given by (3-6) is evaluated using $c(y_T)$, whereas the spatial frequency $f'_y$ given by (3-7) is evaluated using $c(y_R)$. The reason for this apparent discrepancy, although subtle, is very important physically. The spatial frequency $f'_x$ and, hence, the propagation vector component in the $X$ direction must remain constant, that is, its value at the center of the transmit array must equal its value at the center of the receive array in order to be consistent with the solution of the Helmholtz wave equation when the speed of sound is a function of depth. However, since the speed of sound is a function of depth, the spatial frequency $f'_y$ and, hence, the propagation vector component in the $Y$ direction is also a function of depth and, as a result, must be evaluated using the speed of sound at the center of the receive array. The minus signs appearing in (3-6) and (3-7) are a result of the fact that the unit vector normal to the surface of the receive planar array, facing the transmit planar array, points in the negative $Z$ direction since both arrays are assumed to be parallel to the $XY$ plane (see Fig. 1). Also note that the model-based phase weights given by (3-8) are not dependent on—and, from a physical point of view, cannot be dependent on—the index $q$ associated with the $Y$ coordinate of a point source element in the transmit array. A minus sign appears in (3-8) since a plus sign appears in the deterministic ocean medium phase function given by (3-3).

IV. Test Cases and Simulation Results

The model-based signal processing algorithm presented in Section III shall now be evaluated by processing the
composite output signal from the FFT beamformer $\tilde{r}(t)$ by a correlator receiver, followed by a magnitude square operation, and finally by a Neyman–Pearson test (see Fig. 3) in order to determine the impact of the model-based algorithm on the probability of detecting various transmitted electrical signals as a function of the input signal-to-noise power ratio at a single element in the receive array for a given probability of false alarm. The composite signal $\tilde{r}(t)$ can be expressed as

$$\tilde{r}(t) = \tilde{y}(t) + \tilde{n}(t)$$

(4-1)

where

$$\tilde{r}(t) = \sum_{m=-(M-1)/2}^{(M-1)/2} \sum_{n=-(N-1)/2}^{(N-1)/2} \tilde{r}'(t, m, n), \quad (4-2)$$

$$\tilde{y}(t) = \sum_{m=-(M-1)/2}^{(M-1)/2} \sum_{n=-(N-1)/2}^{(N-1)/2} \tilde{y}'(t, m, n), \quad (4-3)$$

and

$$\tilde{n}(t) = \sum_{m=-(M-1)/2}^{(M-1)/2} \sum_{n=-(N-1)/2}^{(N-1)/2} \tilde{n}'(t, m, n). \quad (4-4)$$

Let the processing waveform $\tilde{g}(t)$, which is depicted in Fig. 3, be equal to a time and frequency shifted replica of the complex envelope of the transmitted electrical signal, that is, let

$$\tilde{g}(t) = \tilde{x}(t - \hat{r}) \exp(\pm j2\pi \hat{\phi}), \quad (4-5)$$

where $\hat{r}$ and $\hat{\phi}$ are estimates of the time delay and Doppler shift, respectively. Next, assume that $\tilde{y}(t)$ and $\tilde{n}(t)$ are zero-mean, statistically independent, Gaussian random processes, and that $\tilde{n}(t)$ is wide-sense stationary, low-pass (band-limited) white noise. Therefore, if the output signal components $\tilde{y}'(t, m, n)$ are identical across the array, that is, if rectangular amplitude weights are used and proper beam steering is done via phase weights so that all output signals are cophased, and if the output noise components $\tilde{n}'(t, m, n)$ are uncorrelated, then the theoretical error performance of the receiver shown in Fig. 3 for a Neyman–Pearson test is given by

$$P_D = P_{FA}^{1/(1 + \text{SNR}_A)} \quad (4-6)$$

where $P_D$ and $P_{FA}$ are the probabilities of detection and false alarm, respectively.

$$\text{SNR}_A = (M \times N) \, \text{SNR} \quad (4-7)$$

is the receiver’s output signal-to-noise power ratio due to processing waveforms from an array of elements where 10 log$_{10}$ $(M \times N)$ dB is the array gain,

$$\text{SNR} = \left| X_N(\tau, \phi) \right|^2 \text{SNR}_{\text{in}} \quad (4-8)$$

is the receiver’s output signal-to-noise power ratio due to processing the waveform from a single element in the array [24], $\text{SNR}_{\text{in}}$ is the input signal-to-noise power ratio at a single element in the array and is assumed to be the same at all elements,

$$X_N(\tau, \phi) = \int_{-\infty}^{\infty} \tilde{x}(t) \tilde{x}^*(t - \tau) \exp(\pm j2\pi \phi t) \, dt / E_x \quad (4-9)$$

is the normalized autoambiguity function of the transmitted complex envelope $\tilde{x}(t)$ where

$$\tau = \hat{r} - r_A \quad (4-10)$$

is the error in estimating the actual time delay $r_A$, and

$$\phi = \phi_A - \hat{\phi} \quad (4-11)$$

is the error in estimating the actual Doppler shift $\phi_A$, and

$$E_x = \int_{-\infty}^{\infty} \left| \tilde{x}(t) \right|^2 \, dt \quad (4-12)$$

is the energy of $\tilde{x}(t)$. Note that $X_N(0, 0) = 1$. The corresponding theoretical decision threshold $\gamma$, which is depicted in Fig. 3, is given by

$$\gamma = (M \times N) N_o E_x \ln(1/P_{FA}) \quad (4-13)$$

where $N_o$ (in watts per hertz or joules) is the level of the power spectral density of the band-limited, white, Gaussian, output noise components $\tilde{n}'(t, m, n)$.

The input signal-to-noise power ratio at a single element in the array $\text{SNR}_{\text{in}}$ appearing in (4-8) was defined as follows:

$$\text{SNR}_{\text{in}} \triangleq \frac{E\left\{ \left| \tilde{y}(t, m, n) \right|^2 \right\}}{E\left\{ \left| \tilde{n}(t, m, n) \right|^2 \right\}} = \frac{\left\{ \left| \tilde{y}(t, m, n) \right|^2 \right\}}{\sigma_n^2(m, n)} \quad (4-14)$$

where it was assumed that the random process $\tilde{y}(t, m, n)$ is ergodic, $\langle \cdot \rangle$ indicates time average, and $\sigma_n^2(m, n)$ is the variance of the zero-mean band-limited, white, Gaussian noise $\tilde{n}(t, m, n)$ at element $(m, n)$. Recall that for the simulation results reported in this paper, the output signal components $\tilde{y}(t, m, n)$ are deterministic. Since the FFT beamformer treats $\tilde{r}(t, m, n)$ and, hence, $\tilde{y}(t, m, n)$ as a periodic signal with fundamental period $T_o$ at each element $(m, n),$

$$\left\{ \left| \tilde{y}(t, m, n) \right|^2 \right\} = \frac{1}{T_o} \int_{-T_o/2}^{T_o/2} \left| \tilde{y}(t, m, n) \right|^2 \, dt$$

$$= \sum_{k=-K}^{K} \left| c(q, m, n) \right|^2 \quad (4-15)$$

Fig. 3. Correlator followed by a magnitude square operation.
where
\[ c(q, m, n) = \hat{y}(q, m, n) / L \] (4-16)
are the Fourier series coefficients of \( y(t, m, n) \),
\[ \hat{y}(q, m, n) = \text{DFT}_L \{ y(l, m, n) \}, \]
\[ q = -(L - 1)/2, \ldots, 0, \ldots, (L - 1)/2, \] (4-17)
and
\[ L = 2K + 1 \] (4-18)
is the total number of samples taken at each element \((m, n)\) in the array at a sampling rate of
\[ f_s = L/T_o \text{ samples/s.} \] (4-19)

Therefore, in order to keep the value of the SNR\(_{in} \) the same at all elements \((m, n)\) in the array, the random numbers from the random number generator, which are supposed to be \( N(0, 1) \) and were used to simulate the zero mean, band-limited, white, Gaussian noise samples \( n(l, m, n) \), were scaled by the standard deviation
\[ \sigma_n(m, n) = \left[ \sum_{q = -K}^{K} \left| \hat{y}(q, m, n) \right|^2 / (L^2 \text{SNR}_{in}) \right]^{1/2} \] (4-20)
where use was made of (4-14) through (4-16), \( \hat{y}(q, m, n) \) is given by (4-17) and is different, in general, from element to element, \( L \) is given by (4-18), and the SNR\(_{in} \) is specified. Since both \( y(t, m, n) \) and \( n(t, m, n) \) are baseband with bandwidth
\[ B = Kf_o \text{ Hz,} \] (4-21)
where \( f_o = 1/T_o \) Hz is the fundamental frequency, the noise variance
\[ \sigma_n^2(m, n) = 2BN_o(m, n) = 2Kf_oN_o(m, n) \] (4-22)
and, as a result,\[ N_o(m, n) = \sigma_n^2(m, n) / (2Kf_o) \] (4-23)
is the power spectral density level of the noise at element \((m, n)\). Therefore, since the computer simulation required the noise variance to change from element to element so that the value of the SNR\(_{in} \) remains the same at all elements, the power spectral density level changes from element to element. As a result, the decision threshold \( \gamma \) actually implemented for computer simulation purposes was
\[ \gamma = N_oE_\varepsilon \ln(1/P_{FA}) \] (4-24)
where
\[ E_\varepsilon = \int_{-T_o/2}^{T_o/2} \left| \hat{x}(t) \right|^2 dt = T_o \sum_{k = -K}^{K} \left| c_k \right|^2 \] (4-25)
since \( \hat{x}(t) \) was represented by a finite Fourier series [see (2-42) through (2-44)],
\[ N_o = \sum_{m = -(M-1)/2}^{(M-1)/2} \sum_{n = -(N-1)/2}^{(N-1)/2} N_o(m, n), \] (4-26)
\( N_o(m, n) \) is given by (4-23), and the probability of false alarm \( P_{FA} \) is specified.

Two different common transmitted electrical signals were used in the various test cases. The first type of signal was a single rectangular-envelope CW pulse given by
\[ x(t) = A \text{rect}(t/T) \cos(2\pi f_c t) \] (4-27)
with corresponding complex envelope
\[ \hat{x}(t) = A \text{rect}(t/T) \] (4-28)
where
\[ \text{rect}(t/T) = \begin{cases} 1, & |t| \leq T/2, \\ 0, & |t| > T/2, \end{cases} \] (4-29)
\( T \) is the pulse length in seconds, and
\[ T = dT_o, \quad 0 < d \leq 1, \] (4-30)
where \( d \) is the duty cycle and \( T_o \) is the fundamental period. Since the Fourier transform of (4-28) is given by
\[ \hat{X}(f) = AT \text{sinc}(fT), \] (4-31)
and the Fourier series coefficients that represent \( \hat{x}(t) \) in the time interval \(|t| \leq T_o/2 \) are given by [see (2-44)]
\[ c_k = \hat{X}(kf_o)/T_o, \] (4-32)
substituting (4-31) and (4-30) into (4-32) yields
\[ c_k = Ad \text{sinc}(kd), \quad k = -K, \ldots, 0, \ldots, K \] (4-33)
where
\[ \text{sinc}(x) = \sin(\pi x)/\pi x \] (4-34)
and \( f_oT_o = 1 \). The parameters of the single rectangular-envelope CW pulse used in the various test cases had the following values:
\[ A = 40, \quad d = 0.5, \quad T_o = 5 \text{ ms}, \quad f_c = 5 \text{ kHz}, \]
\[ K = 5, \quad f_{\text{max}} = 6 \text{ kHz} \]
\[ c_0 = 20 \]
\[ c_{-1} = c_1 = 12.73240 \]
\[ c_{-2} = c_2 = 0 \]
\[ c_{-3} = c_3 = 4.244132 \exp(+j180^\circ) \]
\[ c_{-4} = c_4 = 0 \]
\[ c_{-5} = c_5 = 2.546479. \] (4-35)

The second type of transmitted electrical signal used in the various test cases was single rectangular-envelope
LFM pulse given by
\[ x(t) = A \text{rect}(t/T) \cos(2\pi f_c t + bt^2) \] (4-36)
with corresponding complex envelope
\[ \hat{x}(t) = A \text{rect}(t/T) \exp(-jbt^2) \] (4-37)
where \text{rect}(t/T) is given by (4-29), \( T \) is given by (4-30), and \( b \) is the phase deviation constant with units of radians per second. Using the method of stationary phase [16], it can be shown that the Fourier transform of (4-37) is given by
\[ \hat{X}(\omega) = A I^\frac{\pi}{|b|} \exp \left\{ \pm j[\text{sgn}(b) \pi/4 - (\pi f^2/2b)] \right\}, |f| \leq |b| T/(2\pi), \] (4-38)
where
\[ \text{sgn}(x) = \begin{cases} +1, & x > 0 \\ 0, & x = 0 \\ -1, & x < 0 \end{cases} \] (4-39)
is the "signum" or "sign" function and
\[ \frac{4\pi}{|b| T^2} \leq 1 \] (4-40)
must be satisfied [25]. Note that \(|b| T/\pi\) is referred to as the swept bandwidth in hertz. Substituting (4-38) and (4-30) into (4-32) yields
\[ c_k \approx \left( A/T_c \right) (\pi/|b|)^{1/2} \exp \left\{ \pm j[\text{sgn}(b) \pi/4 - (\pi f^2/2b)] \right\}, |k| \leq |b| T_c^2/(2\pi) = K \] (4-41)
which are the Fourier series coefficients used to represent (4-37) in the time interval \(|t| \leq T_o/2\). The parameters of the single rectangular-envelope LFM pulse used in the various test cases had the following values:
\[ A = 40, \quad d = 0.8, \quad T_o = 100 \text{ ms}, \]
\[ f_c = 5 \text{ kHz}, \quad b = 2356.194 \text{ rad/s}^2 \]
\[ K = 3, \quad |b| T/\pi = 60 \text{ Hz}, \quad f_{\text{max}} = 5.03 \text{ kHz} \]
\[ c_o = 14.60593 \exp\left(+j45^\circ\right) \]
\[ c_{-1} = 14.60593 \exp\left(+j21^\circ\right) \]
\[ c_{-2} = 14.60593 \exp\left(+j309^\circ\right) \]
\[ c_{-3} = 14.60593 \exp\left(+j189^\circ\right). \] (4-42)
The following set of parameters was used in the various test cases to establish the relative orientation between the transmit and receive arrays depicted in Fig. 1, and to characterize the arrays themselves.

**Transmit Array:**
\[ x_T = 0 \text{ m}, \quad y_T = 1000 \text{ m}, \quad z_T = 0 \text{ m} \]
\[ M' = N' = 11 \]
\[ d_x = d_y = \lambda_{\text{min}}/2 = 0.1229167 \text{ m}. \] (4-43)

**Receive Array:**
\[ x_R = 500 \text{ m}, \quad y_R = 2500 \text{ m}, \quad z_R = 2549.508 \text{ m} \]
\[ M = N = 5 \]
\[ d_x = d_y = \lambda_{\text{min}}/2 = 0.1229167 \text{ m}. \] (4-44)
The line of sight range \( R_{\text{LOS}} \), that is, the distance along the line from the center of the transmit array to the center of the receive array, was
\[ R_{\text{LOS}} = \left[ (x_R - x_T)^2 + (y_R - y_T)^2 + (z_R - z_T)^2 \right]^{1/2} = 3000 \text{ m}. \] (4-45)
The line of sight angle \( \beta_{\text{LOS}} \), as measured from the positive \( Y \) axis of the transmit array to the line of sight, was
\[ \beta_{\text{LOS}} = \cos^{-1}\left[ (y_R - y_T)/R_{\text{LOS}} \right] = 60^\circ. \] (4-46)
Both the transmit and receive arrays were amplitude weighted by a rectangular amplitude window. The far-field beam pattern of the transmit array was steered toward the center of the receive array.

All the various test cases about to be discussed were separated into two main categories, namely, the homogeneous and inhomogeneous medium test cases identified as HMG1 and INHMG1, respectively. For the various homogeneous medium test cases, the speed of sound \( c \) was constant and was set equal to 1475 m/s. A linear sound-speed profile with constant gradient \( g \) [see (2-11), (2-13), and (2-26)] was used for the various inhomogeneous medium test cases with the speed of sound at the center of the transmit array
\[ c(\gamma_T) = 1475 \text{ m/s and } g = 0.017 \text{ s}^{-1}. \] (4-47)
Two different estimates of time delay \( \check{\tau} \) were used in the processing waveform given by (4-5) for the various test cases. The first estimate was based on the line of sight range \( R_{\text{LOS}} \) and a constant speed of sound for a homogeneous medium, that is,
\[ \check{\tau} = R_{\text{LOS}}/c = 3000/1475 = 2.033898 \text{ s}. \] (4-48)
The second estimate was based on ray acoustics for an inhomogeneous medium [26], that is,
\[ \check{\tau} = \frac{1}{g} \ln \left\{ \tan \left[ \beta(\gamma_R)/2 \right] \right\} \] (4-49)
where \( \beta(\gamma_R) \) is the angle of transmission at the center of the transmit array, and \( \beta(\gamma_T) \) is the angle of arrival at the center of the receive array which can be obtained from Snell’s law as follows [26]:
\[ \beta(\gamma_R) = \sin^{-1}\left\{ \left[ c(\gamma_R)/c(\gamma_T) \right] \sin \beta(\gamma_T) \right\}. \] (4-50)
where
\[ c(y_R) = c(y_T) + g(y_R - y_T). \]  \hspace{1cm} (4-51)

With \( y_T = 1000 \text{ m}, y_R = 2500 \text{ m}, g = 0.017 \text{ s}^{-1}, c(y_T) = 1475 \text{ m/s}, \) and \( \beta(y_T) = \hat{\beta}_{LOS} = 60^\circ, \)
\[ c(y_R) = 1500.5 \text{ m/s}, \beta(y_R) = 61.763^\circ, \]  \hspace{1cm} (4-52)

and
\[ \hat{\tau} = 2.071889 \text{ s}. \]  \hspace{1cm} (4-53)

Since both the transmit and receive arrays were not in motion for the simulation results reported in this paper, \( \phi_0 = 0 \) and, as a result, the estimate of the Doppler shift \( \hat{\phi} \) used in the processing waveform given by (4-5) for the various test cases was
\[ \hat{\phi} = 0 \text{ Hz}. \]  \hspace{1cm} (4-54)

We are now in a position to discuss the computer simulation results of the various test cases obtained by Blount [27]. Note the logical variables “STEER” and “DMEDIA” appearing in the legends of Figs. 4–19. If STEER = TRUE, then standard phase weighting only was done (using (3-4) and (3-5) with the model-based phase weights \( \phi_{MD}(f, n) \) given by (3-8) set equal to zero). If DMEDIA = TRUE, then model-based phase weighting was also done [using (3-8)]. The dashed curves appearing
in Figs. 4-19 represent the theoretical probability of detection that can be obtained when all of the output electrical signals from each element in the receive array are exactly cophased, and when the estimates of the actual time delay and Doppler shift are exact. The dashed curves were computed from (4-6) through (4-8) with the magnitude square of the normalized autoambiguity function set equal to unity in (4-8). The values of probability of false alarm \( P_{fa} \) used in the various test cases were 0.1 and 0.01. The solid curves appearing in Figs. 4-19 represent...
the computer simulation results for the probability of detection ($P_D$) obtained by making relative frequency calculations according to [27]

\[ P_D = \text{HITS/TRIALS}. \tag{4-55} \]

For each value of input signal-to-noise power ratio ($\text{SNR}_{in}$), the simulation was run 100 times for a $P_{FA} = 0.1$ and 500 times for a $P_{FA} = 0.01$ [27].

Figs. 4-7 represent the simulation results for the various homogeneous medium test cases using $\hat{r}$ for a homogeneous medium as given by (4-48). These results represent the baseline performance of the computer simulation. As can be seen from Figs. 4-7; when STEER = TRUE, the simulation agrees reasonably well with theory. For a homogeneous medium, DMEDIA = FALSE.

Figs. 8-15 represent the simulation results for the various inhomogeneous medium test cases using $\hat{r}$ for a homogeneous medium as given by (4-48). Figs. 8-11 for the single CW pulse and, Figs. 12-15 for the single LFM pulse, show the dramatic increase in receiver performance that was obtained when standard plus model-based phase weights were used. The net effect of the model-based phase weights is to force the inhomogeneous medium to act like a homogeneous medium. That is why $\hat{r}$ for a homogeneous medium was used in the correlator receiver. Table I summarizes the approximate increases in receiver performance $\Delta\text{SNR}_{in}$ illustrated by Figs. 8-15. The in-
crease in receiver performance $\Delta SNR_{in}$ was defined as the difference between the $SNR_{in}$ values corresponding to a $P_D = 0.5$ value on the solid curves for (STEER = TRUE, DMEDIA = FALSE) and (STEER = TRUE, DMEDIA = TRUE).

Figs. 16–19 represent the simulation results for the various inhomogeneous medium test cases using standard phase weights only and $\hat{\tau}$ for an inhomogeneous medium as given by (4-53). The increase in receiver performance $\Delta SNR_{in}$ for a $P_D = 0.5$ obtained by comparing the solid curves in Figs. 9 and 16, and Figs. 11 and 17 for the single CW pulse, was approximately 2 dB in both cases (see Table II). However, by comparing Figs. 13 and 18, and Figs. 15 and 19 for the single LFM pulse, it can be seen that standard plus model-based phase weighting using $\hat{\tau}$ for a homogeneous medium results in vastly superior receiver performance (>20 dB) compared to standard phase weighting only using $\hat{\tau}$ for an inhomogeneous medium (see Table II).

And finally, Table III summarizes the approximate increases in receiver performance $\Delta SNR_{in}$ obtained by using standard phase weights only and $\hat{\tau}$ for an inhomogeneous medium. The increase in receiver performance $\Delta SNR_{in}$ for a $P_D = 0.5$ obtained by comparing the solid curves in Figs. 8 and 16, and Figs. 10 and 17 for the single CW pulse, were approximately 15 and 16 dB, respectively. Figs. 12 and 18, and Figs. 14 and 19 for the single LFM pulse, could not be compared directly in order to determine numerical values for receiver performance. However, by comparing these figures, it is clear that receiver performance decreased.

V. SUMMARY AND CONCLUSIONS

The computer simulation results presented in Section IV demonstrated that using standard plus model-based phase weights in conjunction with an FFT beamformer can increase the performance of a correlator receiver dramatically when trying to detect transmitted signals that have propagated through an inhomogeneous medium, and have been corrupted by zero mean, white, Gaussian noise at the receive array. The increase in receiver performance was from approximately 2 to 17 dB and greater depending on the test case (see Tables I and II).

Table I summarizes the increases in receiver performance based on standard plus model-based phase weighting (beamsteering) at the receive array and using $\hat{\tau}$ for a homogeneous medium in the correlator receiver compared to standard phase weighting only and using $\hat{\tau}$ for a homogeneous medium. Since the operation of beamsteering precedes and is independent of the correlator receiver structure, one must first do correct beamsteering in order to achieve the maximum possible array gain. Then, in addition, a correct time-delay estimate is required in the correlator receiver in order to maximize the normalized autoambiguity function, and, hence, the probability of detection. The normalized autoambiguity function decreases in value from its maximum of unity as the errors in estimating time delay and Doppler shift increase. Table I indicates that both waveform types benefit from model-based phase weighting.

Table II summarizes the increases in receiver performance based on standard plus model-based phase weighting (beamsteering) at the receive array and using $\hat{\tau}$ for a homogeneous medium in the correlator receiver compared to standard phase weighting only but using $\hat{\tau}$ for an inhomogeneous medium. Table II indicates that the single LFM pulse benefits most from model-based phase weighting.

Table III summarizes the receiver performance based on standard phase weighting (beamsteering) only at the receive array and using $\hat{\tau}$ for an inhomogeneous medium in the correlator receiver compared to standard phase weighting only and using $\hat{\tau}$ for a homogeneous medium. By comparing the data in Table III to the data in Table I, a decrease in receiver performance of approximately 1 dB was obtained for the single CW pulse when model-based phase weighting was not used. However, a significant decrease in receiver performance was obtained for the single LFM pulse when model-based phase weighting was not used.

Therefore, the results from Tables I–III indicate that
while both waveform types benefitted from model-based phase weighting, the single LFM pulse benefitted the most.

Although the results presented in this paper are based on a limited number of test cases, they do suggest that the concept of model-based signal processing, wherein the detailed physics of a problem is incorporated into signal processing algorithms, has definite merit and warrants further study.

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Ziomek AND Blount: UNDERWATER ACOUSTIC MODEL-BASED SIGNAL PROCESSING 1683


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