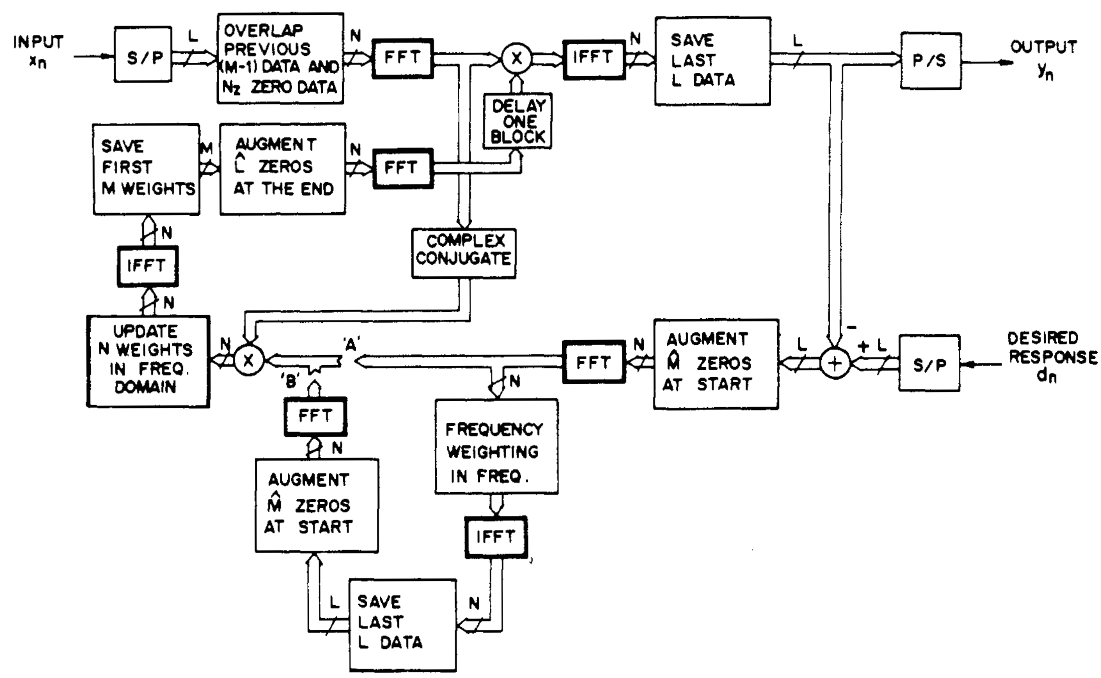

Fig. 1. Realization of FBLMS and FWBLMS ADF's using FFT and overlap–save sectioning procedure ($N = L + M - 1 + N_z$, $\hat{L} \triangleq L - 1 + N_z$, $\hat{M} \triangleq M - 1 + N_z$, and $N_z \geq 0$). [Note: FBLMS ADF is realized with the position "A" connected and FWBLMS ADF is realized with the position "B" connected. N_z is the number of zero data needed for augmenting the input data, thereby allowing to choose a suitable transform of length N. S/P = serial-to-parallel conversion and P/S = parallel-to-serial conversion.]

A Frequency-Weighted Block LMS Algorithm and Its Application to Speech Processing

J. C. LEE, C. K. UN, AND D. H. CHO

In this letter, we propose a frequency-weighted block LMS (FWBLMS) algorithm that is based on minimization of the frequency-weighted block mean-squared error. The FWBLMS algorithm has an important advantage over the time-domain LMS algorithm in that it can be designed with different frequency weighting on error signal depending on the relative significance of various frequency bands. Application of the FWBLMS algorithm in adaptive linear prediction of speech is discussed.

Introduction

Recently, considerable research effort has been directed towards efficient realization of adaptive digital filters (ADFs) using the block least mean square (BLMS) algorithms [1]–[4]. In the BLMS ADFs, filter weights are updated once in every block rather than at every sampling instant. Since the block-estimated gradient becomes cross-correlation between the input and the error signal in the block, the BLMS ADF can be realized efficiently using the fast Fourier transform (FFT). The frequency-domain BLMS (FBLMS) ADF in which filter weights are adjusted in the frequency domain converges fast even for a highly correlated input [1], [2]. In this letter, we investigate another attractive feature of the FBLMS ADF. Specifically, we study a frequency-weighted BLMS (FWBLMS) weight adjustment algorithm that is based on minimizing a weighted sum of frequency-domain errors, and discuss the use of the algorithm in linear prediction of speech for improving the subjective quality of speech.

Manuscript received July 30, 1984; revised October 23, 1984.

The authors are with the Communications Research Laboratory, Department of Electrical Engineering, Korea Advanced Institute of Science and Technology, Seoul, Korea.

A Frequency-Weighted BLMS Algorithm

Let us consider a transversal ADF operated on the block-by-block basis. Prior to derivation of the FWBLMS algorithm, we briefly review the FBLMS ADF. Let M, L, and N be the number of filter weights, the block length, and the transform length of FFT, respectively. The FBLMS ADF shown in Fig. 1 can be obtained by minimizing the frequency-domain block mean-squared error (BMSE). In this ADF, the frequency-domain error vector e_k in the kth block is given by [1]

$$\boldsymbol{e}_k = \boldsymbol{d}_k - \mathscr{P}_{0,L}(\boldsymbol{X}_k \mathscr{P}_{M,0} \boldsymbol{w}_k) \tag{1}$$

where $\boldsymbol{d}_k$ and $\boldsymbol{w}_k$ are the $(N \times 1)$ desired response and filter weight vectors, respectively, both in the frequency domain, and $\boldsymbol{X}_k$ is an $(N \times N)$ diagonal matrix whose diagonal elements are the transformed input data [see Fig. 1]. In (1), the two $(N \times N)$ matrices $\mathscr{P}_{0,L}$ and $\mathscr{P}_{M,0}$ realize the sectioning procedures needed for computing the filter output and adjusting the filter weights, respectively. They are defined as

$$\mathscr{P}_{0,L} \triangleq F \begin{bmatrix} 0 & 0 \\ 0 & I_L \end{bmatrix} F^{-1}$$

and

$$\mathscr{P}_{M,0} \triangleq F \begin{bmatrix} I_M & 0 \\ 0 & 0 \end{bmatrix} F^{-1} \tag{2}$$

where F is the $(N \times N)$ discrete Fourier transform matrix, I_L and I_M denote $(L \times L)$ and $(M \times M)$ identity matrices, respectively, and $\boldsymbol{0}$ is a zero matrix.

As a performance criterion in adjusting the filter weights, the frequency-weighted BMSE (FWBMSE) ε^{fw} is defined by

$$\varepsilon^{fw} \triangleq E\left[\boldsymbol{e}_k^* \Gamma \boldsymbol{e}_k\right] \tag{3}$$

where the asterisk and $E[\cdot]$ denote complex-conjugate transpose of a matrix and statistical expectation, respectively. In (3), Γ is an $(N \times N)$ diagonal matrix whose diagonal elements are of nonnegative values and their relative magnitudes represent the relative

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significance of each frequency component. Following the same approach used for the FBLMS ADF [2], we can have from (1) and (3) a gradient of the FWBMSE with respect to $\boldsymbol{w}_k$ as [5]

$$\nabla \varepsilon^{fw}(\boldsymbol{w}_k) \triangleq \frac{\partial \varepsilon^{fw}}{\partial \boldsymbol{w}_k} = -2E\left[\mathscr{P}_{M,0} X_k^* \mathscr{P}_{0,L} \Gamma \boldsymbol{e}_k\right]. \tag{4}$$

Thus, using an instantaneously estimated gradient, we obtain from (4) an FWBLMS weight adjustment algorithm as the following:

$$\boldsymbol{w}_{k+1} = \boldsymbol{w}_k + \mu \mathscr{P}_{M,0} X_k^* \mathscr{P}_{0,L} \Gamma \boldsymbol{e}_k \tag{5}$$

where μ is a convergence factor controlling the convergence behavior of the algorithm. When the initial values of the filter weights are zero, the algorithm of (5) can be realized alternatively as [1], [5]

$$\boldsymbol{w}_{k+1} = \mathscr{P}_{M,0}\left(\boldsymbol{w}_k + \mu X_k^* \mathscr{P}_{0,L} \Gamma \boldsymbol{e}_k\right). \tag{6}$$

In Fig. 1, a block diagram of the FWBLMS ADF using the algorithm of (6) is shown together with that of the FBLMS ADF. It is noted here that, when Γ is an identity matrix, the FWBLMS algorithm becomes identical to the FBLMS algorithm since $\mathscr{P}_{0,L}\boldsymbol{e}_k = \boldsymbol{e}_k$. Also, it is noted that, when L is sufficiently larger than M, $\mathscr{P}_{0,L}$ can be approximated as an identity matrix. In that case, one can eliminate the FFT and inverse FFT operations that are needed just after the frequency weighting operation in the FWBLMS ADF [see (6) and Fig. 1].

Application of the FWBLMS Algorithm

It is known that some frequency weighting on the error signal in a speech processing system is desirable in improving the subjective quality of speech [6]. One of the commonly accepted aspects of speech perception is that noise in the frequency range of 1 to 3 kHz introduces more serious degradation in intelligibility than that in other range [7].

To investigate the effect of frequency weighting, we have studied adaptive linear prediction of speech using the FWBLMS algorithm by computer simulation, in which we have used a one-step predictor with ten prediction coefficients. The real speech input having unity signal power was sampled at 8 kHz and processed at every 22.5 ms. For frequency weighting we used a C-message weighting function [8] that emphasizes the error signal in the frequency band of 1 to 3 kHz. In Fig. 2, the spectral envelopes of the Hamming-windowed prediction errors of a voiced speech block are compared

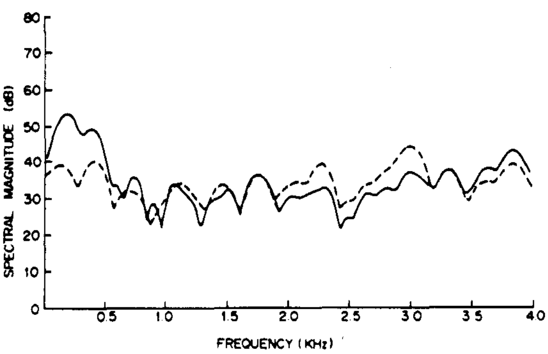

Fig. 2. Spectral envelopes of prediction errors for a voiced speech block with (solid line) and without (dotted line) C-message frequency weighting.

for the cases with and without frequency weighting. It is noted that our simulation results for the cases with and without the $\mathscr{P}_{0,L}$ operation after frequency weighting do not show appreciable differences in the performance of our application example with $M = 10$ and $L = 180$. It is clearly seen from Fig. 2 that, by having more emphasis in the range of 1 to 3 kHz, the frequency-domain error in that range can be reduced at the expense in other frequency region. According to our preliminary results of computer simulation, one can improve the performance of the FWBLMS adaptive linear predictor by more than 2 dB in the frequency-weighted signal-to-noise ratio.

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The Observation of Individual Natural-Frequency Resonances of Radar Targets Through the Scattering of Long Pulses

W. E. HOWELL AND H. ÜBERALL

The Singularity Expansion Method has shown the transient echo amplitude of radar targets to be characterized by complex-frequency poles and their residues. We note that the use of pulsed signals of long duration permits identification of individual pole resonances by the observation of their ringing.

The Singularity Expansion Method (SEM) [1] is based on the observation that the echoes of radar pulses scattered from a finite target appear as the superposition of damped sinusoids,

$$f_{sc}(t) = \sum_{\alpha=1}^{N} R_\alpha e^{\hat{s}_\alpha t} \tag{1a}$$

with complex amplitudes R_α and exponents $\hat{s}_\alpha$. Such a signal shape indicates the presence of complex-frequency poles in the scattering amplitude, via the Laplace transform

$$f_{sc}(s) = \sum_{\alpha=1}^{N} \frac{R_\alpha}{s - \hat{s}_\alpha} \tag{1b}$$

where $s = \omega/i$, and $\omega_\alpha = \operatorname{Im} \hat{s}_\alpha$ are the natural frequencies of the target whose existence gives rise to resonances in the scattering amplitude of (1b).

This latter equation corresponds to a single-frequency excitation of the resonances. If the incident amplitude is pulsed in time, its Fourier spectrum $G(k)$ weighs the factor R_α in (1b), where $k = \omega/c$.

A short pulse is characterized by a wide spectrum $G(k)$; for $G(k) \equiv 1$, one has $f_{inc}(z,t) = \delta(z - ct)$. In such a case, many pole terms contribute to (1b).

An incident pulse of long duration has a narrow spectrum $G(k)$, whose weight in (1b) can radically limit the number of poles contributing to (1). In fact, if the pulse duration is chosen long

Manuscript received June 14, 1984; revised October 17, 1984. Portions of this work were supported by the Naval Research Laboratory.

The authors are with the Department of Physics, The Catholic University of America, Washington, DC 20064, USA.

0018-9219/85/0600-1138$01.00 ©1985 IEEE