An Optimal Gain Control Scheme for Digital On-Channel Repeater with Feedback Interference Canceller in Digital Terrestrial Television Broadcasting Networks

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Abstract: - In this paper, a new gain control scheme is proposed for a digital on-channel repeater (DOCR) with a feedback interference canceller (FIC) in Digital Terrestrial Television Broadcasting (DTTB) networks. In the proposed scheme, a variable attenuator is placed behind the amplifier of the DOCR and the least mean square (LMS) adaptive algorithm is implemented in the FIC to estimate the feedback channel. The proposed scheme consists of two stages. In the first stage, the variable attenuator is fixed on a high attenuation level for a sufficient time with the purpose of avoiding oscillation at start-up of the DOCR. In the second stage, based on the non-oscillation condition derived in Section II, the attenuation level can be decreased at intervals until the aimed gain is approached. Simulations show that the proposed scheme increases the convergence speed of the adaptive algorithm and saves the adjusting time of approaching the aimed gain compared with the conventional gain control scheme.

Key-Words: - DTTB, digital on-channel repeater (DOCR), feedback interference canceller (FIC), oscillation, gain control, least mean square (LMS) algorithm

1 Introduction

In recent years, many countries and regions have published their own DTTB (Digital Terrestrial Television Broadcasting) standards such as DVB-T (Digital Video Broadcasting-Terrestrial) in Europe [1], ATSC (Advanced Television System Committee) in North American [2] and T-DMB (Terrestrial-Digital Multimedia Broadcasting) in South Korea [3]. In these DTTB systems, single frequency network (SFN) which makes efficient use of the spectrum and gets better coverage can be an appropriate choice as an effective and economical approach for DTTB coverage [4, 5].

In order to fill gaps or extend coverage, service providers use a number of digital on-channel repeaters (DOCRs) [6] in SFNs. The DOCR receives DTTB signal from main transmitters, processes it and then retransmits it on the same frequency as received. Although the reception performance can be improved in this way, parts of the retransmitted signal which interfere with the DTTB signal from main transmitters are reflected through the feedback channel and enter the receive antenna. Particularly, if the DOCR's gain is lager than the isolation between the receive antenna and transmit antenna, the power of the feedback interference signal increases continuously through a closed loop composed of the DOCR and the feedback channel and eventually causes the DOCR oscillation.

In general, feedback interference cancellers (FICs) are implemented to prevent DOCRs from oscillating. The FIC estimates the unwanted feedback channel and feedback interference, subtracts the feedback interference estimation from the system input signal and guarantees high isolation between two antennas. Furthermore, adaptive filtering techniques are commonly used in FICs to assist in estimating the unwanted feedback channel [6, 7]. The equalization DOCR (EDOCR) proposed in [8] applies an equalizer to compensate for the multi-path distortion and cancel the feedback interference, but if the power level of the feedback interference is higher than that of the received DTTB signal, the EDOCR can not operate as usual. To solve this problem, an FIC is inserted into the EDOCR for improving the feedback interference cancellation performance [9]. However, the insufficient isolation between two antennas induces the risk of causing the system to start oscillating at startup of DOCRs. To address this early oscillation problem, one way in [10, 11] is sending only the training sequence for quick acquisition of the feedback channel at start-up. Another way in [12, 13] sets the DOCR's gain to a low level in the early stage with the purpose of avoiding oscillation. During the second stage, the gain is increased gradually until the aimed amplification is reached. However, [12, 13] does not derive the gain control scheme specifically.

In this paper, we propose a new gain control scheme by placing a variable attenuator behind the amplifier of the DOCR with an FIC. Meanwhile, the LMS adaptive algorithm is used to estimate the feedback channel and cancel the feedback interference. It is the extension to recent works in [18]. The proposed scheme consists of two stages. In the early stage, the variable attenuator is fixed on a high attenuation level for a sufficient time, which allows the estimation of the feedback channel and the obvious decrease of the coupling between two antennas. And then, based on the non-oscillation condition derived in Section II, the attenuation level can be decreased at intervals until the aimed gain is approached. Subsequently, the LMS algorithm converges and the DOCR operates normally, having attained the aimed gain and interference cancellation levels.

The paper is organized as follows. Section 2 introduces the DOCR system model along with an FIC and derives the non-oscillation condition. In section 3, the conventional gain control scheme is described and the proposed gain control scheme is introduced in section 4. In section 5, the results of simulations support the analysis and discussions. Conclusions are given in section 6.

2 Model of OCR with Echo Canceller

The DOCR system model is shown in Fig. 1. The feedback channel can be modeled as a finite impulse response (FIR) filter.

$$\mathbf{h} = [h_0,\ h_2, \cdots, h_{L-1}]^T \tag{1}$$

where L is the channel impulse response length and $[\bullet]^T$ denotes transpose.

Assuming that the knowledge of the channel impulse response length is known, the channel estimation can be given by

$$\mathbf{w}(n) = [w_0(n),\ w_1(n), \cdots, w_{L-1}(n)]^T \tag{2}$$

where n is the time index .

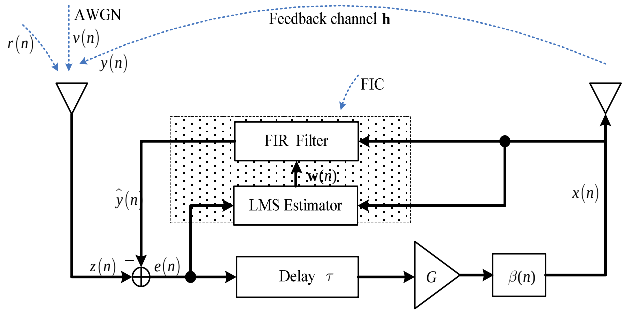

Fig.1 DOCR system model

In the single frequency mode of DTTB networks, the received signal of DOCRs consists of the target signal coming from the main transmitter, feedback interference and local noise signal

$$z(n) = r(n) + \mathbf{h}^H \mathbf{x}(n) + v(n) \tag{3}$$

where $\mathbf{x}(n) = [x(n),\ x(n-1), \cdots, x(n-L+1)]^T$, $x(n)$ is the reference signal of the FIR filter with zero mean and variance $\sigma_x^2 = E|x(n)|^2$, $[\bullet]^H$ denotes conjugate transpose. For simplicity, assume that $v(n)$ is additive white Gaussian noise with zero mean and variance $\sigma_v^2 = E|v(n)|^2$ and $r(n)$, uncorrelated with $v(n)$, is zero mean and variance $\sigma_r^2 = E|r(n)|^2$.

The FIC subtracts the feedback interference estimation at the receiving end. Thereby, we get the error signal

$$\begin{aligned} e(n) &= z(n) - \mathbf{w}^H(n)\mathbf{x}(n) \\ &= r(n) + \mathbf{h}^H \mathbf{x}(n) - \mathbf{w}^H(n)\mathbf{x}(n) + v(n) \\ &= r(n) + \Delta\mathbf{w}^H(n)\mathbf{x}(n) + v(n) \end{aligned} \tag{4}$$

where $\Delta\mathbf{w}(n) = \mathbf{h} - \mathbf{w}(n)$ denotes the weight error vector which represents the residual feedback interference channel.

Using the LMS algorithm [13], the weight update equation operated in the channel estimator is

$$\mathbf{w}(n+1) = \mathbf{w}(n) + 2\mu e^*(n)\mathbf{x}(n) \tag{5}$$

where μ is the step-size parameter.

The output signal $x(n)$ is given by

$$x(n) = G \bullet \beta(n) \bullet e(n-\tau) \tag{6}$$

where G is the DOCR's gain, $\beta(n)$ represents the attenuation value of the variable attenuator at time n and τ is the processing delay time of the DOCR which is considered to be an integer times of the sampling interval.

Substituting (4) into (6), $x(n)$ can be written by

$$\begin{aligned} x(n) = {} & G \bullet \beta(n) \bullet r(n-\tau) + \\ & G \bullet \beta(n) \bullet \Delta\mathbf{w}^H(n)\mathbf{x}(n-\tau) \\ & + G \bullet \beta(n) \bullet v(n-\tau) \end{aligned} \tag{7}$$

The requirement of the variable attenuator at time n is derived in Appendix as

$$\beta^2(n) < \frac{1}{G^2 \bullet E\left\{\left\|\Delta \mathbf{w}(n)\right\|^2\right\}} \quad (8)$$

where $E\left\{\left\|\Delta \mathbf{w}(n)\right\|^2\right\} = \sum_{i=0}^{L-1} E\left\{\left|\Delta w_i(n)\right|^2\right\}$ which represents the mean square deviation (MSD) [14], $\|\bullet\|$ is the Euclidean norm of a vector.

This condition prevents oscillation and guarantees finite DOCR output power, and it is ensured by the gain control scheme in this paper.

3 Conventional Gain Control Scheme

In order to prevent the system from oscillating at startup, two stages in the gain control process are established: the early stage and the follow-up stage [12]. The early stage operates only once during the initialization or startup of the DOCR. During this stage, the gain stays low for a sufficient time to allow the convergence of the adaptive algorithm and the reduction of the coupling, so that it is stable independently of the cancellation level reached by the adaptive filter. In the follow-up stage, the gain increases slowly until the aimed gain and allowed maximum output power are approached. Subsequently, the DOCR performs normally, having attained the aimed gain and cancellation levels. Note that the function of adjusting the DOCR's gain can also be carried out by adjusting the variable attenuator which is placed behind the amplifier. In order to facilitate the comparison between two gain control schemes in this paper, the adjusting process can be described by

$$\beta^2(n+1) = \begin{cases} \beta^2_{\min}, & \text{if } n < k_1 N \\ \beta^2(n), & \text{if } n \ge k_1 N, \text{and } \mathrm{mod}(n,N) \neq 0 \\ \min\left\{\beta^2(n) + \Delta_{\text{fix}},\ \beta^2_{\max}\right\} & \text{if } n \ge k_1 N, \text{and } \mathrm{mod}(n,N) = 0 \end{cases} \quad (9)$$

where k_1 and N are integer, N is the period of attenuator adjustment which represents the attenuator adjustment is only performed at every N samples. Δ_{fix} is a pre-defined value which represents the fixed adjusting value of the attenuator at every N samples. In order to avoid oscillation, Δ_{fix} can not be pre-defined too large. $\beta^2_{\min}$ is the minimum attenuation value and $\beta^2_{\max}$ is the aimed maximum attenuation value, $\mathrm{mod}(\bullet,\bullet)$ is the modulus function. In practice, $\beta^2_{\min} < \beta^2_{\max} \le 1$.

We denotes the method described (9) in this section as the conventional gain control (CGC) method.

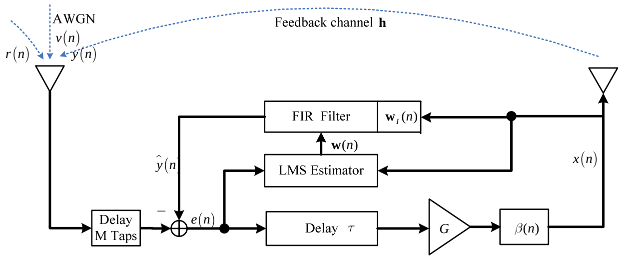

Fig.2 proposed gain control scheme

4 Proposed Gain Control Scheme

Based on the non-oscillation condition (8), we derive a new gain control scheme. Generally, in order to preserve the waveform quality of the received target signal from the main transmitter, a sufficient, pre-defined gain margin β_0^2 which is typically taken to be 10dB to 15dB is required [15]. According to (8), the allowed maximum value of the variable attenuator at time n is given by

$$\beta^2(n) = \frac{1}{G^2}\frac{1}{\beta_0^2}\frac{1}{E\left\{\left\|\Delta \mathbf{w}(n)\right\|^2\right\}} \quad (10)$$

Note that $\beta^2(n)$ depends on the MSD. In the initial state for DOCRs, the MSD may remain to be a large value due to the insufficient convergence of the LMS algorithm. Thus, $\beta^2(n)$ preserves a high attenuation level. After a while, with the decrease of the MSD, $\beta^2(n)$ increases gradually until the aimed gain is approached. Additionally, if the MSD increases due to a small variation of the feedback channel, $\beta^2(n)$ decreases to make high attenuation and avoid oscillation.

Due to the unknown feedback channel, the MSD in (10) is not accessible. In order to obtain the square deviation estimation, an artificial delay of M samples is inserted into the signal path of $Z(n)$ before the sub-traction yielding $e(n)$, as shown in Fig. 2. The output signal is artificially delayed by M samples. Furthermore, the FIR filter is increased by M taps, resulting in an extended FIR filter $\mathbf{w}_e(n)$, that is

$$\mathbf{w}_e(n) = \begin{bmatrix} \mathbf{w}_I(n) \\ \mathbf{w}(n) \end{bmatrix} \quad (11)$$

where $\mathbf{w}_I(n) = [w_{I,0}(n),\ w_{I,1}(n), \cdots, w_{I,M\text{-}1}(n)]^T$.

The optimal values of the inserted weight coefficients corresponding to the artificial delay are zero. We may utilize the known property that the LMS algorithm tends to spread the weight-error evenly among the filter coefficients [14, 16]. Thus, using the norm of the inserted weight coefficients, we may approximate the MSD as

$$\widehat{D}_{\mathrm{w}}(n) \approx \frac{L}{M}\sum_{k=0}^{M-1}\left|w_{I,k}(n)\right|^2 = \frac{L}{M}\left\|\mathbf{w}_I(n)\right\|^2 \tag{12}$$

The extension of M taps for the FIR filter increases memory requirement of M elements. For every iteration of the LMS algorithm, the complexity is increased by M complex multiplications and additions. Appropriate choice of M can limit the complexity increase.

According to (10) and (12), the allowed maximum value of the variable attenuator at time n is given by

$$\widehat{\beta}^2(n) = \frac{1}{G^2}\frac{1}{\beta_0^2}\frac{1}{\widehat{D}_{\mathrm{w}}(n)} \tag{13}$$

Therefore, the adjustment for the variable attenuator operates in the following manner

$$\beta^2(n+1) = \begin{cases} \beta_{\min}^2, & \text{if } n < k_1 N \\ \beta^2(n), & \text{if } n \geq k_1 N, \text{and} \\ & \qquad \mathrm{mod}(n,N) \neq 0 \\ \min\left\{\beta^2(n)+\Delta(n) \quad \beta_{\max}^2\right\} & \text{if } n \geq k_1 N, \text{and} \\ & \qquad \mathrm{mod}(n,N) = 0 \end{cases} \tag{14}$$

where $\Delta(n)$ is the adjusting value of the attenuator at time n whose bound is

$$0 < \Delta(n) \leq \Delta_{\max}(n) \tag{15}$$

and $\Delta_{\max}(n)$ is the coupling decrease during the period of N samples which limits the adjusting value at time n and is given by

$$\begin{aligned}\Delta_{\max}(n) &= \widehat{\beta}^2(n) - \widehat{\beta}^2(n-N) \\ &= \frac{1}{G^2}\frac{1}{\beta_0^2}\left[\frac{1}{\widehat{D}_{\mathrm{w}}(n)} - \frac{1}{\widehat{D}_{\mathrm{w}}(n-N)}\right]\end{aligned} \tag{16}$$

The behavior of the gain control scheme (16) can be described as follows: the variable attenuator is fixed on a high attenuation level for a sufficient time (a number of $k_1 N$ consecutive samples) in the early stage. In it, the LMS algorithm estimates the feedback channel and reduces the coupling to the desired level. And then, based on the estimation of the MSD, the variable attenuation level can be decreased at intervals until the aimed gain is approached.

5 Simulations

5.1 Simulation Setup

In this section simulation examples are presented to evaluate the performance of the two gain control methods. All figures are obtained by averaging the results over 100 trials of the same experiment. The experimental OFDM signal used in the simulations is generated based on Chinese DTTB standard PN420+C3780 mode [17]. The duration of one signal frame is 555.6μs in this mode. The symbol rate specified in the standard is 7.56 Mbps. However, we use a square root raised cosine (SRRC) filter with a roll-off factor of 5% as shaping filter to limit the bandwidth of the transmitted signal to 8MHz. The sampling rate in simulations is 16MHz. The power amplifier gain is chose to 42dB and the isolation between the antennas is set to -14dB. The processing delay of the DOCR is assumed to be 10μs. The adjusting period N is chose to 8400 samples which are the samples of one frame DTTB signal and k_1 is set to 10. We choose the maximum attenuation value $\beta_{\max}^2 = 0\text{dB}$ and the minimum value $\beta_{\min}^2 = -40\text{dB}$, i.e., the pre-defined gain margin β_0^2 is 12dB in simulations which satisfies the gain margin discussed in Section VI. The length of the FIR filter is 20 taps and the artificial delay is 10 taps.

Two channels are involved in the simulations. The channel between the main transmitter and the DOCR which is modeled as Brazil channel A with 30dB signal to noise ratio (SNR) at the DOCR's receiving end. Another is the feedback channel between the receive antenna and the transmit antenna. Power delay profile of each channel is presented in Table 1.

Table.1 Parameters of Channels

Brazil channel A		Feedback channel	
Relative Power (dB)	Delay (μs)	Relative Power (dB)	Delay (μs)
0	0	0	0
-13.8	0.15	-3	0.32
-16.2	2.22	-8	0.62
-14.9	3.05	-15	1.2

5.2 Definitions

The normalized square deviation (NSD) is a measure of the convergence properties of the LMS algorithm with the gain control schemes, which is defined as

$$\mathrm{NSD}(n) = 10\log_{10}\left(\frac{\left\|\mathbf{w}(n)-\mathbf{h}\right\|^2}{\left\|\mathbf{h}\right\|^2}\right) \tag{17}$$

Note that if the lengths of $\mathbf{w}(n)$ and $\mathbf{h}$ differ, the shorter one is padded with zeros when computing the NSD.

The echo return loss enhancement (ERLE) is used to show the feedback interference cancellation performance, defined as

$$\mathrm{ERLE}(n) = 10\log_{10}\left(\frac{\sum_{k=0}^{K-1}\left|z(n-k)\right|^2}{\sum_{k=0}^{K-1}\left|e(n-k)\right|^2}\right) \tag{18}$$

where K is the length of a window over which the ERLE is computed and is set to 8400.

5.3 Results

The adjusting process of the variable attenuator for the CGC and the proposed gain control scheme is shown in Fig. 3. We can see that in the proposed gain control scheme the attenuator spends only about 9.5ms (about 17 frame signal) for approaching the aimed gain and allowed maximum output power, compared to 27.5ms (about 50 frame signal) in the CGC scheme with $\Delta_{\text{fix}} = 1\text{dB}$ and 16.5ms (about 30 frame signal) with $\Delta_{\text{fix}} = 2\text{dB}$. Δ_{fix} can not be pre-defined too large. In the simulation, if Δ_{fix} is larger than 3dB, the system may come into oscillation. This situation indicates that setting Δ_{fix} too large may lead to dissatisfaction for the non-oscillation condition (8) whereas the proposed scheme which is derived from (8) can avoid this situation. In addition, the attenuator adjusting value in the proposed scheme is decreased gradually with the frame number, this is attributed to the convergence property of the LMS algorithm. In the initial stage of adaptation, the large output error leads to the high convergence speed and fast coupling decrease between two antennas. When the adaptive filter goes toward the steady-state, the output error becomes small and the decrement of the coupling is reduced. In Fig. 3, we can observe that the last adjusting value is about 2.5dB before approaching the aimed gain.

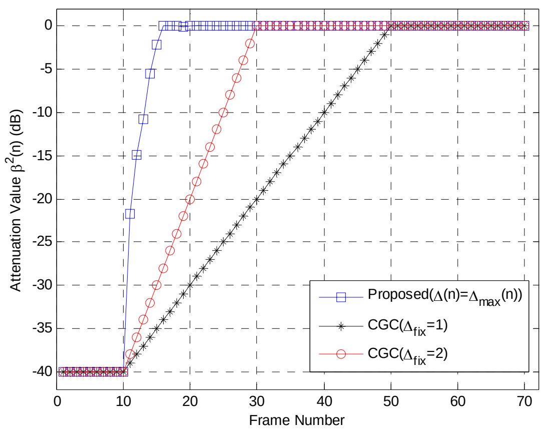

Fig.3 Attenuation value comparison for the CGC and proposed gain control scheme

Fig. 4 shows the adjusting performance for different values of M. Setting M too low may increase the adjusting time for approaching the aimed gain, this is because too low M may give rise to influence on the accuracy of the MSD estimation. On the other hand, the complexity increases with an increased value of M. So the choice of M is a tradeoff and it depends on the application at hand.

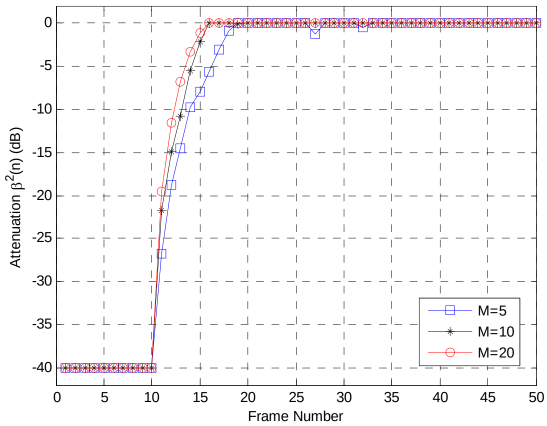

Fig.4 Attenuation value comparison for different setting of M in the proposed gain control scheme

Fig. 5 shows the learning curves of NSD for the CGC and proposed schemes. The learning curve of NSD is averaged with a rectangular smoothing window of 8400 samples. The NSD curve of the proposed scheme converges faster which can be attributed to the attenuator adjustment in manner of (14). We can see that in the proposed gain control scheme the adaptive algorithm spends only about 11ms (about 20 frame signal) for approaching the steady-state, while the CGC scheme spends about 22ms(about 40 frame signal) with and $\Delta_{\text{fix}} = 2\text{dB}$ and 33ms(about 60 frame signal) with $\Delta_{\text{fix}} = 1\text{dB}$ for converging. The improvement of the convergence rate for the adaptive algorithm may lead to the faster approaching of the steady state of the DOCR and the decrease of the residual feedback interference in the gain adjusting process. Fig. 6 shows the learning curve of the ERLE performance for the CGC and proposed scheme.

In Fig. 6, about 28dB feedback interference is cancelled by the FIC either in the CGC or in the proposed scheme. However, it is apparent that the time of approaching the optimal ERLE value spent in the proposed scheme is shorter than that spent in the CGC scheme, this is due to the shorter adjusting time of approaching the aimed gain in the proposed scheme.

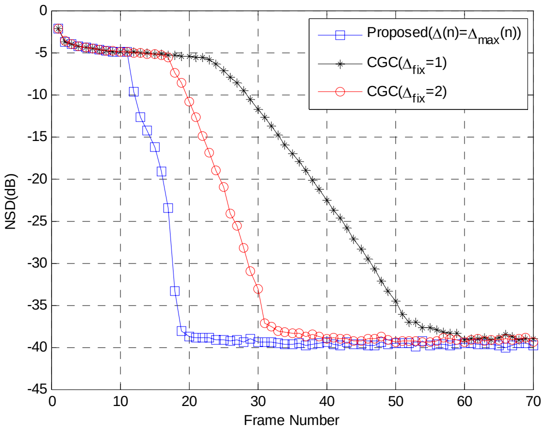

Fig. 5 NSD performance of the feedback channel estimation for the CGC and proposed gain control scheme

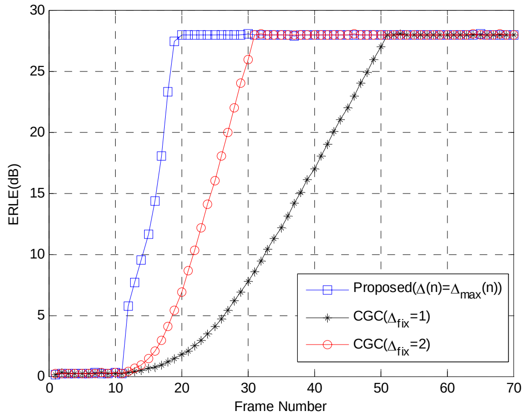

Fig.6 ERLE performance for the CGC and proposed gain control scheme

6 Conclusion

In this paper, a new gain control scheme based on the non-oscillation condition for a DOCR with an FIC is proposed. Compared with the CGC scheme, the proposed scheme focuses on avoiding oscillation at startup of DOCRs and decreasing the adjusting time for approaching the maximum output power. Simulation results support the theoretical analysis and discussions. In addition, except for the Chinese DTTB networks, this scheme is proper to anywhere in an environment which needs DOCRs, such as DVB-T, ATSC and DMB-T networks.

Appendix
Derivation of (6)

The output signal power is calculated from (7) by

$$\begin{aligned}E\left\{x^2(n)\right\} &= G^2\beta^2(n)\sigma_r^2 \\ &+G^2\beta^2(n)E\left\{\Delta\mathbf{w}^H(n)\mathbf{x}(n-\tau)\mathbf{x}^H(n-\tau)\Delta\mathbf{w}(n)\right\} \\ &+G^2\beta^2(n)\sigma_v^2\end{aligned} \quad (19)$$

With assumption that the amplified output signal of DOCRs is uncorrelated with the weight error vector [14] and the attenuator remains invariable for an observation period, the second term on the right side of (19) can be expressed by

$$\begin{aligned}&G^2\beta^2(n)E\left\{\Delta\mathbf{w}^H(n)R(n-\tau)\Delta\mathbf{w}(n)\right\} \\ &= G^2\beta^2(n)\mathrm{tr}\left\{E[R(n-\tau)\Delta\mathbf{w}(n)\Delta\mathbf{w}^H(n)]\right\} \\ &= G^2\beta^2(n)\sum_{i=0}^{L-1}\sum_{j=0}^{L-1}E\left\{\Delta w_i(n)x(n-i-\tau)x^*(n-j-\tau)\Delta w_j^*(n)\right\} \\ &= G^2\beta^2(n)\sum_{i=0}^{L-1}\sum_{j=0}^{L-1}E\left\{x(n-i-\tau)x^*(n-j-\tau)\right\}E\left\{\Delta w_i(n)\Delta w_j^*(n)\right\} \\ &= G^2\beta^2(n)\sum_{i=0}^{L-1}E\left\{\left|x(n-i-\tau)\right|^2\right\}E\left\{\left|\Delta w_i(n)\right|^2\right\} \\ &+G^2\beta^2(n)\sum_{\substack{i=0\\ i\neq j}}^{L-1}\sum_{j=0}^{L-1}E\left\{x(n-i-\tau)x^*(n-j-\tau)\right\}E\left\{\Delta w_i(n)\Delta w_j^*(n)\right\} \\ &= G^2\beta^2(n)\sigma_x^2\sum_{i=0}^{L-1}E\left\{\left|\Delta w_i(n)\right|^2\right\}+T(n)\end{aligned} \quad (20)$$

where

$$R(n-\tau)=E\left\{\mathbf{x}(n-\tau)\mathbf{x}^H(n-\tau)\right\} \quad (21)$$

$$T(n)=G^2\beta^2(n)\sum_{\substack{i=0\\ i\neq j}}^{L-1}\sum_{j=0}^{L-1}E\left\{x(n-i-\tau)x^*(n-j-\tau)\right\}E\left\{\Delta w_i(n)\Delta w_j^*(n)\right\} \quad (22)$$

and $\mathrm{tr}(\bullet)$ is the trace operator, $T(n)$ is the cross-correlation term. Under the usual independence assumption that the successive increments of tap weights are independent of one another [19], $T(n)$ can be seen as a small value in general.

Substituting (20) into (19), we get

$$\sigma_x^2=\frac{G^2\beta^2(n)(\sigma_r^2+\sigma_v^2)+T(n)}{1-G^2\beta^2(n)\sum_{i=0}^{L-1}E\left\{\left|\Delta w_i(n)\right|^2\right\}} \quad (23)$$

where the non-oscillation condition is

$$G^2\beta^2(n)\sum_{i=0}^{L-1}E\left\{\left|\Delta w_i(n)\right|^2\right\}<1 \quad (24)$$

This means the power of the feedback interference is attenuated lower than that of the target signal through the closed loop composed of the FIC, the amplifier, the variable attenuator and the feedback channel.

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