order model in the sense of making an error norm based upon energy arbitrarily small. Moreover, the optimal reduction depends upon specification of a time interval; and if the system is nonlinear will also in general depend upon the particular initial conditions. If the original system contains a driving function, the reduced order model will depend on that as well.

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On the Implementation of Adaptive Electronic Hybrid for Digital Subscriber Loops

CHE-HO WEI AND NENG-AN KUO

Abstract **—For digital transmission over the existing two-wire subscriber lines with echo cancellers, the capability of echo suppression is an important consideration for bidirectional transmission. To reduce the burden of the echo canceller, the adaptive electronic hybrid is proposed to improve the transhybrid loss. In this paper, we first briefly review the theory of the adaptive electronic hybrid. A simple hardware implementation with linear integrated circuits is described. The hybrid is tested under simulated line conditions. Experimental results show that realized adaptive electronic hybrid can achieve at least 26-dB transhybrid loss.**

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I. Introduction

For the future Integrated Service Digital Network (ISDN), the CCITT has recommended 144 kbps $2B+D$ channels for subscriber lines. To implement the digital transmission over the existing two-wire subscriber loops, echo-cancellation method is one of the most promising methods proposed [1]. The amplitude attenuation of the transmitted signal from far-end transmitter to near-end receiver could be as high as 50 dB. However, the conventional fixed hybrid can only achieve about 11-dB transhybrid loss with 3 dB standard deviation [2]. To receive the far-end signal correctly, it is required that the amplitude ratio between the received far-end signal and feedthrough echo of the near-end signal must be at least 20 dB. Normally, the echo canceller was added to the fixed balancing hybrid to gain 60 dB attenuation of undesirable feedthrough echoes. Due to the inherent nonlinearities of A/D and D/A converters in echo canceller, 60 dB attenuation of echo is difficult to achieve [3]. To release the burden of high degree suppression of echo, an adaptive electronic hybrid is used to increase the transhybrid loss.

In this paper, we first briefly review the theory of adaptive electronic hybrid. Hardware implementation of the adaptive electronic hybrid by analog integrated circuits is given to demonstrate its capability of echo suppression.

II. Analysis of Adaptive Electronic Hybrid

In transmission system, the hybrids provide the connection between two- and four-wire ports. In the conventional hybrid, the balancing network is fixed, regardless of which subscriber line is connected. To improve the separation capability of the hybrid, the adaptive technique must be utilized in the design of hybrid circuit. Recently, Korn [4] has developed a model for adaptive hybrid for digital subscriber lines. A block diagram of the adaptive electronic hybrid connected with the subscriber line is shown in Fig. 1. The echo-suppression adaptive filter $C(s)$ produces an estimated $\hat{e}(t)$ such that the undesirable $e(t)$ is cancelled at the end of the receiver. In Fig. 1, $v_F(t)$ is the voltage transmitted from the far-end of the line, Z_1 and Z_2 are the terminating impedances on near-end and far-end of the line, respectively. The voltage $v_L(t)$ is an attenuated version of $v_F(t)$ at the receiver end. The near-end received signal can be expressed as

$$V_R(s)=V_T(s)[H(s)-C(s)]+V_L(s) \tag{1}$$

where $H(s)$ is the transfer function of the echo path from $v_T(t)$ to $e(t)$. When $C(s)=H(s)$, we obtain $V_R(s)=V_L(s)$. Hence $C(s)=H(s)$ is the object of echo-suppression adaptive filter.

The subscriber line can be represented by its primary constants $(R_\mu, L_\mu, C_\mu, G_\mu)$ or secondary constants (characteristic impedance Z_0, propagation constant γ), where R_μ, L_μ, C_μ, and G_μ are resistance, inductance, capacitance and conductance per unit length of the transmission line. The input–output relationship can be described by the $ABCD$ transmission matrix as [4]

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} A & B \\ C & D \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \tag{2}$$

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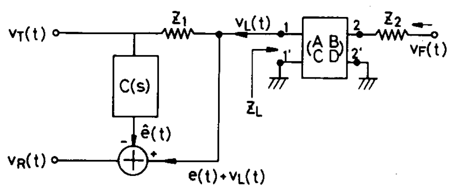


Fig. 1. Model of adaptive electronic hybrid connected with subscriber line.

where A, B, C, D are given by

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} \cosh(\gamma l) & Z_0 \sinh(\gamma l) \\ \sinh(\gamma l)/Z_0 & \cosh(\gamma l) \end{pmatrix} \tag{3}$$

and $\gamma = (R + j\omega L\mu)(G + j\omega C\mu)$. From (3), the input impedance viewed from port 1 is given by

$$Z_L = \frac{V_1}{I_1} = \frac{AV_2 + BI_2}{CV_2 + DI_2} = \frac{AZ_2 + B}{CZ_2 + D}. \tag{4}$$

Let $v_F = 0$, then the echo path transfer function is obtained as

$$H(s) = \frac{V_1(s)}{V_T(s)} = \frac{Z_L}{Z_1 + Z_L} = \frac{AZ_2 + B}{CZ_1Z_2 + DZ_1 + AZ_2 + B}. \tag{5}$$

According to (2), by setting $V_T = 0$ we obtain the transfer function of the channel as

$$H_F(s) = \frac{V_1(s)}{V_F(s)} = \frac{(AD - BC)Z_1}{CZ_1Z_2 + DZ_1 + AZ_2 + B}. \tag{6}$$

The transfer function $H(s)$, from (5), can be rearranged as

$$H(s) = \frac{AR + B}{CR^2 + DR + AR + B}. \tag{7}$$

Let us assume that (7) can be approximated by an equivalent rational function [5]

$$\tilde{H}(s) = r_0 + \sum_{i=1}^{L} \frac{r_i}{s + p_i} \tag{8}$$

with r_i, p_i being real constants. Consequently, we apply a similar rational function to approximate $C(s)$, i.e.,

$$\tilde{C}(s) = \hat{r}_0 + \sum_{i=1}^{L} \frac{\hat{r}_i}{s + \hat{p}_i} \tag{9}$$

with $\hat{r}_i, \hat{p}_i$ being the estimated versions of r_i and p_i, respectively.

One criterion for achieving optimum matching is to minimize the mean-squared value $E[v_R^2(t)]$ (dc power). Assuming that $v_T(t)$ and $v_F(t)$ are uncorrelated stationary random processes, we then obtain $Q(\hat{r}_i, \hat{p}_i) = E(v_R^2(t)] = E[d^2(t)] + E[v_L^2(t)]$, where $d(t) = e(t) - \hat{e}(t)$. The well-known least mean-square algorithm [6] gives the following relation [5]:

$$\frac{d\hat{x}_i}{dt} = -2K_i v_R(t) \frac{\partial v_R(t)}{\partial \hat{x}_i} \tag{10}$$

or

$$\hat{x}_i = -2K_i \int v_R(t) \frac{\partial v_R(t)}{\partial \hat{x}_i} dt$$

where $\hat{x}_i$ stands for every $\hat{r}_i$ and $\hat{p}_i$, and K_i is the integration gain. Equation (10) is an unbiased case of the steepest descent algorithm and its functional block diagram is shown in Fig. 2. To compute $\partial v_R(t)/\partial \hat{x}_i$, we first take the inverse Laplace transform in both sides of (1) and then take the partial derivative with respect to $\hat{x}_i$. Thus we have the relation

$$\frac{\partial v_R(t)}{\partial \hat{x}_i} = \int_S -V_T(s) \frac{\partial C(s)}{\partial \hat{x}_i} e^{st}\, ds. \tag{11}$$

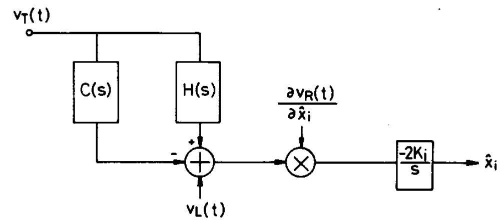


Fig. 2. Block diagram of steepest descent algorithm.

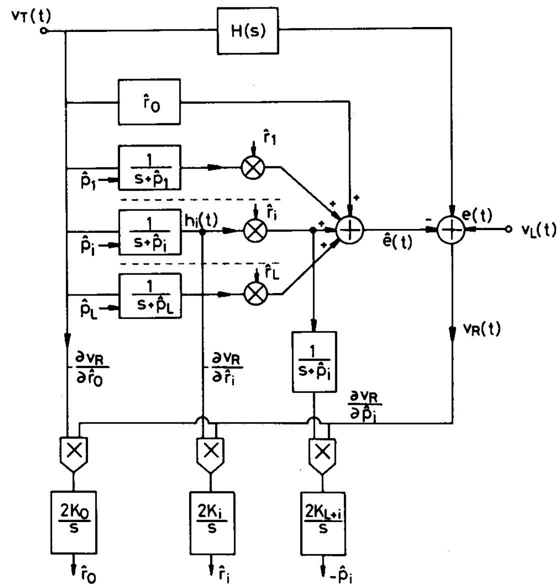


Fig. 3. Block diagram of adaptive electronic hybrid.

From (11), we can obtain

$$\frac{\partial V_R(s)}{\partial \hat{r}_0} = -V_T(s)$$

$$\frac{\partial V_R(s)}{\partial \hat{r}_i} = -\frac{1}{s + \hat{p}_i} V_T(s) = -h_i(s)$$

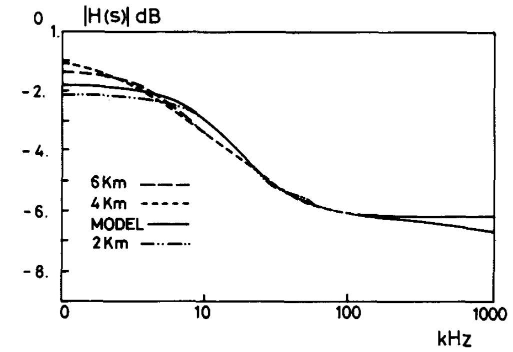
$$\frac{\partial V_R(s)}{\partial \hat{p}_i} = \frac{\hat{r}_i}{(s + \hat{p}_i)^2} V_T(s) = f_i(s). \tag{12}$$

With (10) and (12), the adaptive algorithm is demonstrated explicitly in Fig. 3.

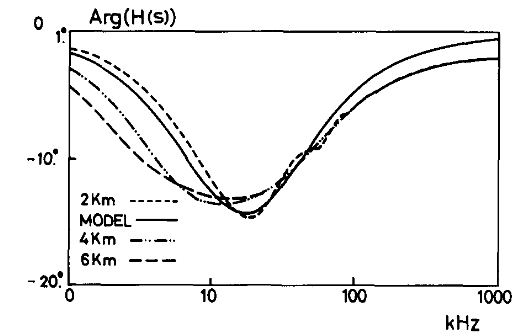
Although the hybrid has the residues $\hat{r}_i$ and poles $\hat{p}_i$ as adjustable parameters, we can adopt the special case that only the residues $\hat{r}_i$ are adjustable but the poles are assumed to be constant. The echo-suppression adaptive filter of this special case is given by

$$\tilde{C}(s) = \hat{r}_0 + \sum_{i=1}^{L} \frac{\hat{r}_i}{s + \hat{p}_{i0}} \tag{13}$$

where $\hat{p}_{i0}$ are some real constants. Note that the same adaptation can be utilized for seeking the optimum residues. Generally, the

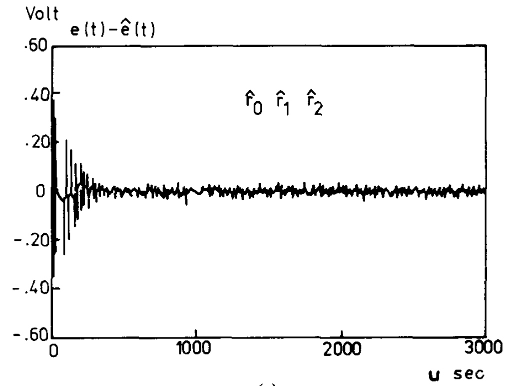


(a)

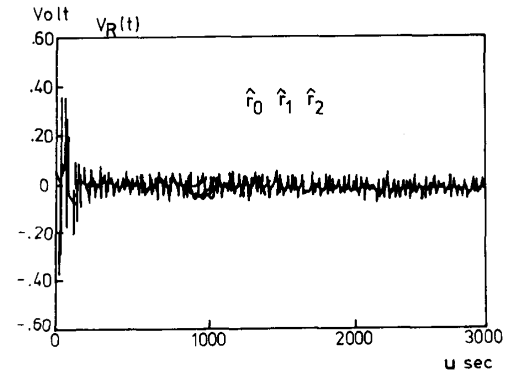


(b)

Fig. 4. (a) Magnitude response of transfer function $H(s)$. (b) Phase response of transfer function $H(s)$.



(a)



(b)

Fig. 5. (a) Convergence behavior of echo residual in three-residue hybrid in the absence of far-end signal. (b) Convergence behavior of received signal $v_R(t)$ when far-end signal is present.

constants $\hat{p}_{i0}$ are determined by computer simulation that estimates the possible distribution of poles.

The magnitude and phase responses of $H(s)$ must be determined from the actual subscriber line simulations. Fig. 4 shows a typical frequency response of $H(s)$ for 2-, 4-, and 6-km 0.5-mm CCPLAP lines. The response of a five-parameter $(r_0, r_1, r_2, p_1, p_2)$ model $\hat{H}(s)$ is also shown in the figure. It is found that

$$\hat{H}(s) \approx 0.48 + \frac{1300}{s+42000} + \frac{1900}{s+210000}.$$

Computer simulation of a three-residue adaptive electronic hybrid (r_0, r_1 and r_2 are adaptive with fixed p_1 and p_2) connected with a 4-km, 0.5-mm CCPLAP line is shown in Fig. 5. About 0.35 ms is required for 35 dB attenuation of residual echo $e(t)-\hat{e}(t)$, in the absence of far-end signal. In general the residual echo is slightly larger when far-end signal is present.

III. Hardware Implementation

The hardware implementation of the adaptive electronic hybrid is straightforward by using linear integrated circuits. The key analog circuit is the adaptive gain amplifier to realize $\hat{r}_i$. The gain of the amplifier is automatically controlled by the output of an integrator. There are two ways to realize the automatic gain control (AGC). The first one is by using voltage-controlled resistor such that a linear control function of gain is achieved with a field-effect transistor (FET). The second one is by the application of a linear multiplier, such as MC1494. The output signal of the multiplier is a linear function of a control signal connected to one of its inputs. The poles, p_1 and p_2 are realized by integrators. The detailed circuit configuration of a three-residue hybrid is given in Fig. 6. The integration gains can be adjusted by varying the values of m_i and R, C components to fit the dynamic range of the processed signal.

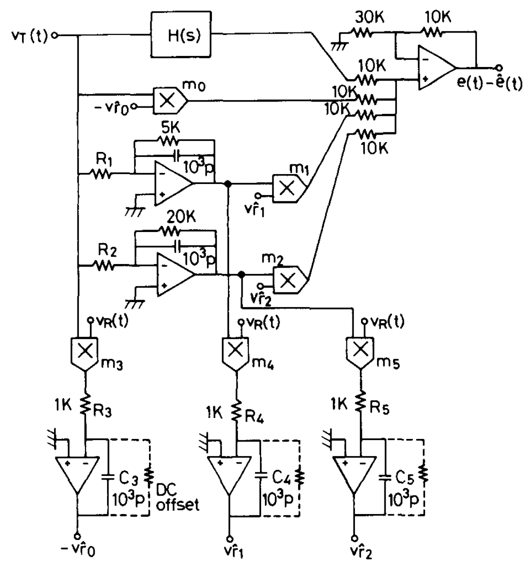
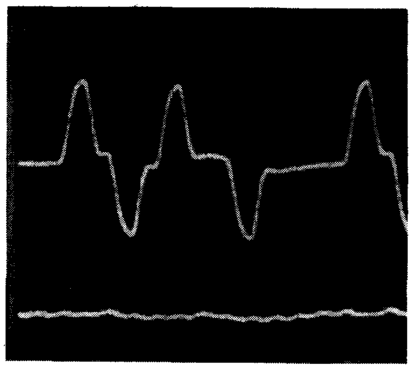
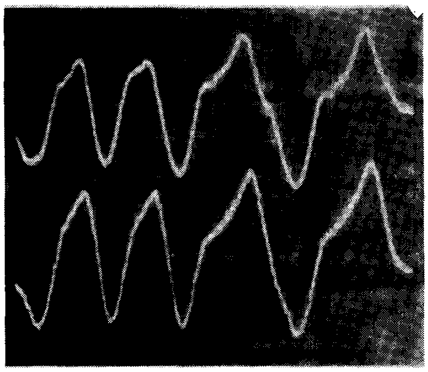


Fig. 6. Circuit diagram of adaptive electronic hybrid with residues as adjustable parameters only.



(a)



(b)

Fig. 7. (a) Waveforms of echo and echo residual in the three-residue hybrid system in the absence of far-end signal. (Top curve: $e(t)$; bottom curve: $e(t)-\hat{e}(t)$. (1 volt/div.). (b) Waveforms of far-end signal and near-end received signal. (Top curve: $v_L(t)$; bottom curve: $v_R(t)$. (0.1 volt/div.).

The hardware is implemented and tested under simulated line conditions. Both near- and far-end signals are encoded by Alternative Mark Inversion (AMI) codes. The experimental results are shown in Fig. 7. In the absence of far-end signal, the adaptive electronic hybrid can achieve about 30-dB transhybrid loss. The echo suppression capability is defined as the ratio between the peak value of $e(t)$ and $v_R(t)$ when no far-end signal is added. In the presence of far-end signal, the adaptive hybrid still works properly, regardless of the level of the far-end signal. Fig. 7(b) shows the added far-end signal $v_L(t)$ and the near-end received signal $v_R(t)$ when a sequence of near-end transmitted signal is transmitted. When the line condition is changed, the experimental results show that the three-residue hybrid still has at least 26 dB transhybrid loss. If the hybrid has more adjustable parameters, it is less sensitive to the variation of line conditions.

IV. Conclusions

Adaptive electronic hybrid is proposed to replace the conventional fixed hybrid in the digital subscriber loop to reduce the burden of echo canceller. The theory of the adaptive hybrid is briefly reviewed. A simple method of hardware implementation is described. In this method, the poles of the echo path transfer function are assumed to be constant. A three-residue adaptive electronic hybrid is realized by linear integrated circuits. Experimental results show that the realized hybrid can achieve at least 26 dB transhybrid loss.

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Spatial-Domain Design of Three Terms Separable Denominator Two-Dimensional Digital Filters

AKIRA TAGUCHI AND NOZOMU HAMADA

***Abstract*—A new type of two-dimensional (2-D) digital filter named three terms separable denominator (3TSD) filter is considered. And the spatial domain design techniques for it are presented. Transfer function of approximated filter is obtained by solving a set of linearized equation of bilinear equations iteratively. The 3TSD filter also has some benefits as SD filter. For examples, SD filter is easy to check its stability and simple to implement. An example is presented to illustrate the utility of the proposed technique.**

I. Introduction

In recent years, two-dimensional (2-D) digital filters have been extensively investigated, especially with applications to image processing. Several design methods of 2-D digital filter have been reported. In this paper, we present a spatial-domain procedure for designing 2-D causal IIR filter with new type of transfer function.

Since design methods of 2-D IIR filter with general form of transfer function are confronted with the difficulties in implementation and stability test, many design techniques have been proposed under the separable restriction on denominator polynomial. Transfer function of 2-D separable denominator (2-D SD) IIR filter is written by

$$H_{SD}(z_1, z_2) = \frac{N(z_1, z_2)}{D_1(z_1) D_2(z_2)}. \quad (1)$$

Thus the stability of this is checked for z_1 and z_2 independently. Also, many established design methods for 1-D filter can be modified to the method for 2-D SD filter design [3], [5], [6], [10], [11].

We now briefly examine the restriction on the frequency response of 2-D SD filter, especially in connection with quadrantally symmetric magnitude response. And, we will introduce new type of transfer function named three terms separable denominator (3TSD), given by

$$H(z_1, z_2) = \frac{N(z_1, z_2)}{D_1(z_1) D_2(z_2) D_3(z_1 z_2)}. \quad (2)$$

This type of filter is expected to provide better approximation

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