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# AN ADAPTIVE MIXED-SIGNAL NARROWBAND INTERFERENCE CANCELLER FOR WIRELINE TRANSMISSION SYSTEMS

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**ABSTRACT**

Narrowband radio transmitters like radio amateurs and broadcast radio stations are considered to be a serious problem for high-bitrate data transmission over twisted pairs. Due to its high power level, radio frequency interference (RFI) has the potential of overloading the receiver's analog-to-digital converter (ADC). Once the ADC is in saturation, any countermeasure taken in digital domain will fail, so the problem has to be faced at least partly in the analog domain. This paper proposes an adaptive, mixed-signal, narrowband interference canceller employing a modified recursive least-squares (RLS) algorithm which is split into an analog and a digital part. The mixed-signal approach enables the circuit to generate an interference-cancelling signal of several MHz while operating the adaptive algorithm at some kHz. The structure is fast enough to prevent the ADC from overloading due to radio amateur interference, thus protecting the data transmission from interruption. Simulation results as well as measurements indicate a practical disturbance rejection potential of about 40 − 50 dB.

## 1. INTRODUCTION

Emerging high-speed data transmission systems intended for use in the access part of the public telephony network, such as the *digital subscriber line* (DSL) family, use a much larger bandwidth than the twisted copper pairs were originally intended for. This introduces impairments that researchers and engineers have not been faced with in voice-band modem technology. Among these, *radio-frequency interference* (RFI) caused by broadcast AM radio transmitters and radio amateurs is considered to be one of the most challenging problems. Especially the interference from radio amateurs (HAM interference) is difficult to handle. It is non-stationary, as the transmission is intermittent and bursty, and exhibits potentially high power levels [1] when transmitters are close to the wiring. The effect of HAM interference could be compared to having somebody shouting in your ear while trying to listen to a polite conversation.

A considerable amount of work has been done to mitigate the RFI problem in the digital domain, *i.e.*, after the analog-to-digital converter (ADC) [2]−[5]. However, none of these approaches work in case the ADC is saturated. This paper presents a novel RFI canceller structure essentially solving the ADC overload problem. The feasibility of our proposed canceller, originally reported in [6], has been verified by means of a prototype.

The paper proceeds as follows. Section 2 provides a system-level overview of the problem and Section 3 introduces the canceller. Simulation results are discussed in Section 4 and the prototype measurement results are presented in Section 5. Section 6 compares the proposed canceller to the canceller described in [7], to our knowledge the only approach to this problem published so far. Finally, Section 7 concludes the work.

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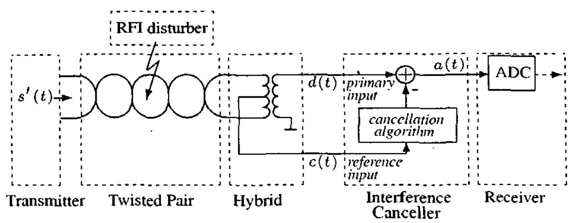


Figure 1: Principle of the RFI canceller. The desired data is carried by the differential mode signal $d(t)$. The common mode signal $c(t)$ provides the reference for the canceller.

## 2. THE RFI PROBLEM

Figure 1 illustrates the system setup. We aim at transmitting the signal $s'(t)$ over a twisted copper pair. The wire pair acts like an antenna and picks up radio signals, especially when installed as an overhead line. This is often the case for the last meters of wire to the subscriber, which unfortunately is also where the HAM transmitter most often is located. Basically the disturbance couples onto the line as a common mode signal. This common mode interference signal can be very strong if the transmitter is close to the wire. Since wires in the field are never perfectly balanced, the common mode disturbance is partially converted into a differential mode component which interferes additively with the signal we would like to receive. This results in a strong differential mode disturbance, but the common mode signal also provides an almost perfect reference.

Let $c(t)$ be the common mode component of the narrowband disturber at the receiver. The resulting differential mode signal $d(t)$ at the receiver input can be written as the sum of the received signal



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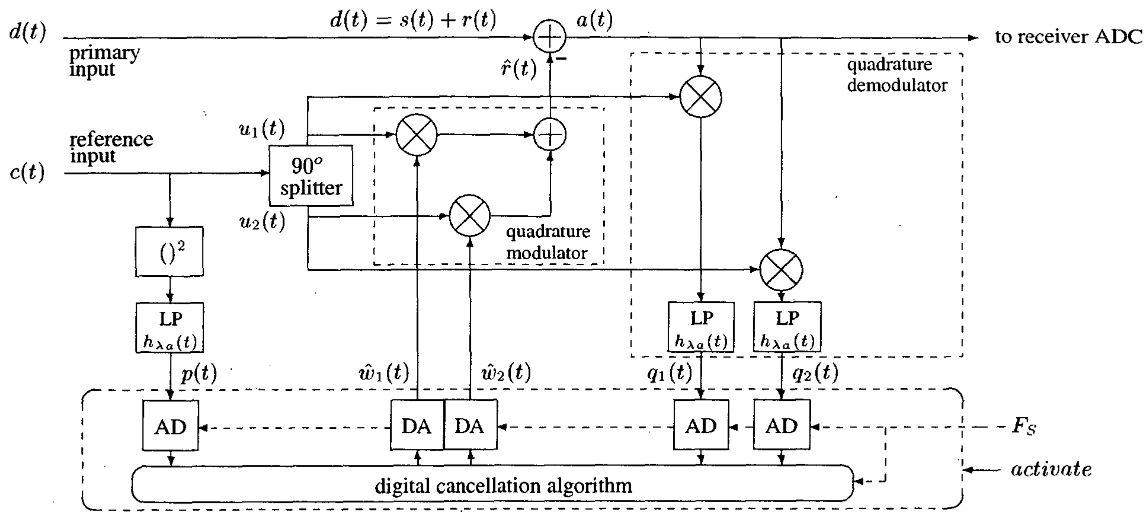


Figure 2: Canceller block diagram.

$s(t)$ and the differential disturber component $r(t)$, *i.e.*,

$$d(t) = s(t) + r(t) = s(t) + a_{c2d}\, c(t + \tau_{lag}), \qquad (1)$$

where $a_{c2d}$ denotes the coupling factor and $\tau_{lag}$ takes into account a small shift in time. In [2], a speed difference between common mode and differential mode signals at 1 MHz of 0.31 times the speed of light is reported. However, multiple coupling between the two modes and the fact that a considerably long part of the wire may be hit by the disturber, complicate the scenario. RFI ingress measurements on a dropwire indicate a lag between $-1\,\mu$s and $+1\,\mu$s in the frequency range up to 30 MHz [8]. Depending on the type of wire, the common mode to differential mode coupling $a_{c2d}$ can be as strong as $-35$ dB. The common mode disturbance can be in the order of 30 $V_{\text{peak}}$ at the receiver, which may result in differential mode disturber levels of up to 0.5 $V_{\text{peak}}$ [9], [10]. The level of the desired signal at the receiver's input (we will refer to this as the far-end signal), for a medium length wire of 1.5 km, is typically in the range of $60-80$ $mV_{\text{peak}}$. The receiver's ADC, tuned to sample only the desired far-end signal, will then saturate, and the desired signal is lost.

An important part of understanding the RFI problem and its countermeasures, lies in understanding the nature of the RFI-signal itself. To be able to combat RFI disturbers, we use a combination of high-frequency analog and low-frequency digital signal processing. It may at first seem contradictory that a canceller can use analog-to-digital and digital-to-analog converters clocked at some kHz when the signal to be cancelled is at several MHz. However, although the carrier frequencies are high, the bandwidth of the signals themselves is only a few kHz. Usually disturbers have a bandwidth between two and three kHz. What we really need to track with the canceller are changes in the coupling from the common mode signal to the differential mode signal, *i.e.*, $a_{c2d}$ and $\tau_{lag}$ in Equation (1). They are both frequency dependent. However, their change is virtually zero within the few kHz of bandwidth of a HAM-disturber. Neither does the disturber traverse along the line at any speeds that would cause rapid changes in the coupling. Thus, we can assume that $a_{c2d}$ and $\tau_{lag}$ are constant.

## 3. CANCELLER STRUCTURE

The fact that the disturbing signal is of special nature, *i.e.*, of small bandwidth, can be exploited when designing a canceller [11]. The differential mode interference $r(t)$ is virtually the same as the reference signal $c(t)$, except that it is scaled by $a_{c2d}$ and shifted in time by $\tau_{lag}$, as given in Equation (1).

### 3.1. Signal decomposition

Figure 2 shows a block diagram of the canceller. The analog-to-digital and digital-to-analog converters are operating at the sampling frequency $F_S$, which is in the range of some (say 20) kHz. Note that the signals at the ADCs are at baseband. As illustrated in Figure 2, the circuit has two inputs: a primary input for the differential mode signal $d(t)$ and a reference input for the common mode signal $c(t)$. Every $T = 1/F_S$ seconds, the cancellation algorithm calculates a new coefficient vector

$$\hat{\boldsymbol{w}}[n] = \begin{bmatrix} \hat{w}_1[n] \\ \hat{w}_2[n] \end{bmatrix}, \qquad (2)$$

which is converted into the time-continuous weight signals $\hat{w}_1(t)$ and $\hat{w}_2(t)$ by the two digital-to-analog converters, *i.e.*,

$$\hat{\boldsymbol{w}}(t) = \begin{bmatrix} \hat{w}_1(t) \\ \hat{w}_2(t) \end{bmatrix} = \hat{\boldsymbol{w}}[n], \quad nT \le t < (n+1)T. \qquad (3)$$

The canceller generates an interference-cancelling signal $\hat{r}(t)$ that yields the residual error estimate

$$\xi(t) = r(t) - \hat{r}(t), \qquad (4)$$

which depends on $t$ and on the current weights. The common mode signal $c(t)$ serves as a reference and is converted into the Hilbert pair

$$\boldsymbol{u}(t) = \begin{bmatrix} u_1(t) \\ u_2(t) \end{bmatrix} \qquad (5)$$

by means of a $90^o$ phase-shift device. The elements of $\boldsymbol{u}(t)$ are weighted by $\hat{\boldsymbol{w}}(t)$ to generate the interference-cancelling signal

$$\hat{r}(t) = \hat{\boldsymbol{w}}(t)^{\mathrm{T}}\boldsymbol{u}(t), \tag{6}$$

where $(\cdot)^{\mathrm{T}}$ represents the transpose. Note that (6) is realized by the quadrature modulator in Figure 2. With the two parameters $\hat{w}_1(t)$ and $\hat{w}_2(t)$, both amplitude and phase of the interference-cancelling signal $\hat{r}(t)$ can be arbitrarily adjusted. The quadrature demodulator generates the baseband error signal $\boldsymbol{q}(t)$, which is the lowpass filtered product of the Hilbert pair $\boldsymbol{u}(t)$ and the error estimate $\xi(t)$ caused by the current weight vector, *i.e.*,

$$\boldsymbol{q}(t) = \begin{bmatrix} q_1(t) \\ q_2(t) \end{bmatrix} = h_{\lambda a}(t) * (\boldsymbol{u}(t)\,\xi(t)), \tag{7}$$

where $h_{\lambda a}(t)$ is the lowpass filter impulse response and $*$ denotes the convolution operator. Furthermore, a measure of the common mode reference signal power

$$p(t) = h_{\lambda a}(t) * c(t)^2 \tag{8}$$

is essential for the cancellation algorithm. The signals $q_1(t)$, $q_2(t)$ and $p(t)$ are sampled at the rate $F_S$ by three ADCs.

### 3.2. Mixed-Signal Realization of the Cancellation Algorithm

The algorithm is based on the following model of the differential mode signal:

$$d(t) = \boldsymbol{w}(t)^{\mathrm{T}}\boldsymbol{u}(t) + s(t). \tag{9}$$

The far-end signal as well as any noise present at the receiver input is represented by $s(t)$. We have to find an update estimate $\hat{\boldsymbol{w}}[n]$ for the weight vector $\boldsymbol{w}[n]$, given the estimate at iteration $n-1$, *i.e.*, $\hat{\boldsymbol{w}}[n-1]$ as well as information contained in the observable signals $\boldsymbol{u}(t)$ and $d(t)$. Note that the development of the classical recursive least-squares (RLS) algorithm [12] is based on similar preliminaries. Hence, we define the cost function

$$\varepsilon[n] = \frac{1}{T}\int_0^{nT} \lambda^{nT-t}|e(t)|^2 dt, \tag{10}$$

where

$$e(t) = r(t) - \hat{\boldsymbol{w}}[n]^{\mathrm{T}}\boldsymbol{u}(t) \tag{11}$$

is the estimation error and the constant $\lambda \leq 1$ is a forgetting factor weighting recent data higher and older data lower. We assume that the lowpass filtered elements of $\boldsymbol{u}(t)$ are uncorrelated, *i.e.*,

$$h_{\lambda a}(t) * (u_1(t)\,u_2(t)) = 0, \tag{12}$$

and have the same power

$$h_{\lambda a}(t) * u_1(t)^2 = h_{\lambda a}(t) * u_2(t)^2. \tag{13}$$

Minimizing (10) yields the update rule for the weight vector

$$\hat{\boldsymbol{w}}[n] = \hat{\boldsymbol{w}}[n-1] + \frac{1}{P[n]}\boldsymbol{q}(nT), \tag{14}$$

with

$$P[n] = \lambda P[n-1] + p(nT). \tag{15}$$

Equations (14) and (15) constitute the digital part of the update algorithm. The analog part comprises the weighting within one period T, carried out by the lowpass filters in (7) and (8). By minimizing (10), we obtain the optimum lowpass filter impulse response

$$h_{\lambda a}(t) = \begin{cases} \frac{1}{T}\lambda^{t/T} & 0 \leq t \leq T \\ 0 & \text{else} \end{cases}. \tag{16}$$

In the current prototype canceller, we approximate this impulse response by a first order passive RC lowpass filter.

## 4. SIMULATION RESULTS

We evaluate the results derived in the previous section by time-domain simulations. Different signal scenarios as well as various canceller parameter settings have been investigated. The simulations results presented in this section have been selected to demonstrate both the performance and the limits of the canceller. The input signals, shown in Figure 3, constitute a challenging scenario, defined by the following parameters: $F_S = 20$ kHz, HAM disturber frequency $= 3.7$ MHz, power of disturber $= 0$ dBm, far-end signal power $= -30$ dBm, $a_{c2d} = -35$ dB. The disturber turns on, goes off, and then returns at the same frequency. The circuit is activated by an external trigger signal that may be derived from the common mode signal power. After waiting one clock-cycle for the analog filters to stabilize, the canceller then puts out the coefficients and immediately removes most of the differential mode disturbing signal. When the disturber disappears and the canceller is deactivated by the trigger signal, the current weights are stored. The second time the disturber switches on, the canceller starts with the correct (stored) weights. Using the weights of the previous activation is reasonable since HAM disturbance is bursty.

The loss of signal to noise ratio due to RFI interference is the most meaningful performance measure, as used *e.g.* in [13]. However, since our main focus is avoiding an ADC overload condition, we identify the RFI suppression, *i.e.*, the power ratio of the residual disturbance at the canceller output to the interference present at the differential mode input, as the measure of interest. Figure 4 shows the simulated RFI suppression level when activating the canceller. In relation to the disturber's power evolution the RFI suppression rises fast enough to protect the ADC from overload.

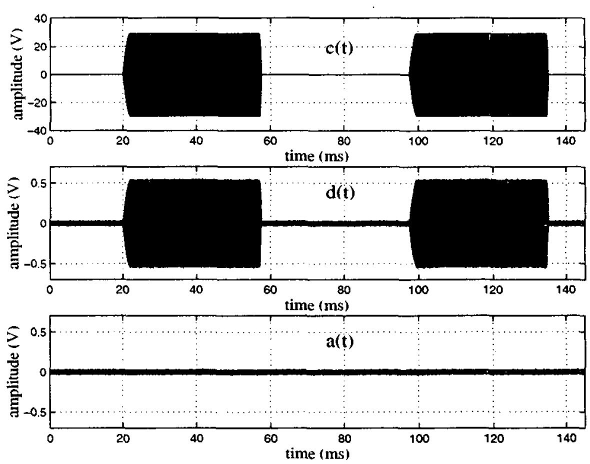


Figure 3: Simulated canceller performance: common mode input signal $c(t)$, differential mode input signal $d(t)$, and canceller output (receiver's ADC input) signal $a(t)$. Note the different voltage scaling for common mode and differential mode signals.

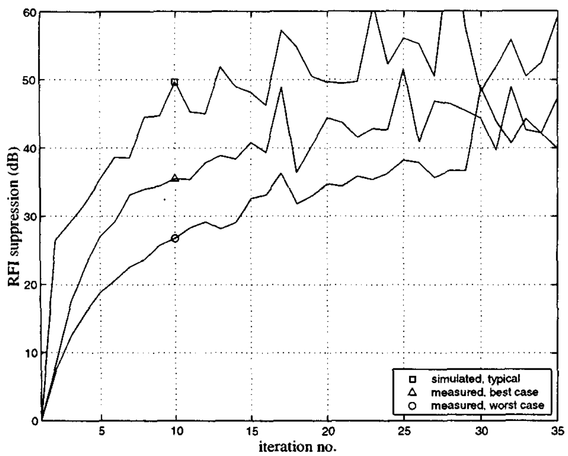


Figure 4: Measured canceller performance compared to simulation: evolution of RFI suppression - best case and worst case of the measurement series and typical simulation result.

## 5. MEASUREMENT RESULTS

A prototype of the canceller has been built using a PC with a 12-bit AD-/DA-card. The algorithm is implemented in software using Delphi. As far as possible, the experiment has been setup according to the parameters summarized in Section 4. However, at the time of writing, the operating frequency $F_S$ of our prototype was limited to 2 kHz due to port access constraints imposed by the AD-/DA-card. Figure 4 shows the measured evolution of the RFI suppression. The curves correspond to best and worst cases out of a series of experiments. As expected, imperfections of the analog circuitry, especially the multipliers, reduce the prototype performance compared to simulation. However, the results confirm the feasibility of our approach. The prototype achieves a steady state RFI suppression of at least 40 dB.

## 6. COMPARISON

We briefly compare the proposed canceller to the canceller of [7], since it is the only approach to the analog RFI problem published so far. However, the integrated solution of [7] is targeted for a time-division duplex system, assuming that there exists a fairly long silent period each data frame. This silent period can be used to run the adaptive algorithm without being disturbed by the signal. During data transmission the coefficients are frozen, *i.e.*, there is no tracking of the interferer. The algorithm in [7] is of least-mean-square (LMS) type and a steady state interference rejection of 30 dB is reported. Although the canceller in [7] works perfectly in the system it was designed for, it cannot be used in our more general application. As far as can be deduced from [7], where RFI cancellation is a side issue, our approach gives faster convergence and at least 10 dB better suppression.

## 7. CONCLUSIONS

RFI cancellation before the receiver's ADC in wireline transmission systems is both necessary and difficult. The HAM disturbance is bursty and exhibits high power levels. The main goal of the analog interference canceller is to prevent the ADC from overloading.

A novel mixed-signal cancellation scheme is proposed. It generates an interference-cancelling signal of several MHz, while the adaptive algorithm operates at a rate of some kHz. This is essentially achieved by splitting the RLS algorithm into an analog and a digital part, where the high-frequency signal processing is analog. Simulation results, supported by prototype measurements, are presented. Once adjusted, the canceller achieves a narrowband interference suppression of about $40-50$ dB. The canceller is fast enough to protect data transmission over copper twisted pairs from HAM radio interference.

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