Abstract—In our Software-Defined-Radio (SDR) project we aim to combine two receivers (HiperLAN/2 and Bluetooth) on one common platform. In this paper, the main focus is on one of the performance bottlenecks of such a receiver, namely the bandpass filter section in Bluetooth mode. Contributions of inter-symbol interference $ISI$, adjacent channel interference $ACI$ and noise on the total bit error rate $BER$ are analyzed. The influence of the channel selection filter characteristics on these contributions are researched. Larger values for (for instance) transition bandwidth result in lower order filters that reduce the computational load, which is an important design consideration. This also reduces bit errors caused by $ISI$, one of the two major contributors to the total $BER$. On the other hand, lower order filters increase the $BER$ due to $ACI$. Optimal filter parameters are derived from these trade-offs and applied the system which will be presented.

I. INTRODUCTION TO THE SDR PROJECT

In our Software-Defined-Radio (SDR) project [4] we aim to combine two receivers (HiperLAN/2 and Bluetooth) on one common platform. Our focus is from antenna output to raw bits. In our opinion, SDR is an implementation technology which aims at providing flexibility and reconfigurability to hardware platforms. The HiperLAN/2 hardware is that complex compared to the Bluetooth hardware, that Bluetooth capability may be added to the HiperLAN/2 platform at limited costs. So, it is not the demand for flexibility (one front end for all signals) that motivates us, but the idea of providing added functionality nearly for free. We've built a test bed with two separate receivers (HiperLAN/2 and Bluetooth), partly in hardware (analog front end) and partly in software (digital front end). This paper will only focus on the digital channel selection part of these two receivers (shown in figure [1]).

The analog front end of our demonstrator (described in [1]) includes two Analog to Digital Converters (ADCs) and produces a complex 80 MSPS signal $z[n]$ (see figure [1]). For HiperLAN/2 the output is one channel at baseband (with a nominal bandwidth of 20 MHz) as our HiperLAN/2 demodulator requires a complex 20-MSPS baseband signal. On the other hand, for Bluetooth, the output of the analog front end is a "chunk" of 20 Bluetooth channels (with a nominal bandwidth $B_{BT}$ of 1 MHz each) [2]. Our Bluetooth demodulator requires a real bandpass signal with center frequency located at 2.5 MHz (with 10 samples per symbol [5]). So the digital channel selection executes three functions: mixing (only Bluetooth), filtering and downconversion.

First an introduction will be given to the channel selection system of the receiver. Both HiperLAN/2 and Bluetooth modes will be described. Then, a more in-depth presentation is given of the Bluetooth channel selection requirements and design considerations. After discussing the $BER$ calculation methods and trade-offs of the system, some experiments are defined to evaluate the system. After discussing the results, conclusions are drawn and final specifications of the current (test bed) system are given.

II. INTRODUCTION TO THE CHANNEL SELECTION SYSTEM

The current channel selection system operates in two different modes, depending on the input signal: HiperLAN/2 or Bluetooth. The input signal $z[n]$ (see
z[n] \rightarrow \text{LPF} \rightarrow \text{M} \rightarrow b[n'] \\
\text{z[n]} \rightarrow \text{LPF} \rightarrow \text{M} \rightarrow b[n']

Fig. 2
HiperLAN/2 digital channel selection system,
\( M = 4 \)

The second step is performed by the first bandpass filter \( BPF1 \). The bandwidth of \( BPF1 \) is defined as \( B_1 \). Assuming the Bluetooth Signal-Of-Interest (SOI) has a carrier frequency \( f_c = 4.5 \) MHz, the filter cutoff frequencies \( f_{1,p}, f_{1,s} \) are defined as \( f_c \pm B_1/2 \). The resulting signal \( e[n'] \) is mixed with \( f_{io} = 2 \) MHz. Output \( f[n'] \) now contains two images of the SOI at \( f'_{c,1}, f'_{c,2} = 2.5, 6.5 \) MHz. The \( BPF2 \) filter removes the unwanted image with (in this case) \( f'_{c,2} = 6.5 \) MHz. After the second decimation by factor \( M2 = 2 \) the signal is ready for demodulation.

A. Bluetooth: requirements

The main objective of this design is to achieve the \( BER \), required under well defined circumstances, with the least amount of computations. For Bluetooth, a maximum allowable \( BER \) is defined for certain worst case input scenarios \( S \). This \( BER \) \( (1 \cdot 10^{-3}) \) will be calculated using simulations of over \( 3 \cdot 10^4 \) bits. For a given demodulator, the \( BER \) performance is directly related to the signal-to-noise ratio \( (SNR_{demod}) \) of the signal \( S \). \( SNR_{demod} \) depends on the received carrier power, analog filter bandwidth \( B \) and the received interference.

The received interference consists of the reception of one other Bluetooth signal. This interfering signal can have four different carrier frequencies \( (f_{co}, f_{a,1}, f_{a,2} \) and \( f_{a,3} \) \) and strengths (listed below, frequencies are given in MHz) \( [S] \):

\[
\begin{align*}
\text{f}_{co} & \quad \text{Co-channel interference with } f_{co} = f_c \text{ power is } -11 \text{ dB relative to the SOI.} \\
\text{f}_{a,1} & \quad \text{First adjacent channel interferer (closest neighbor) with } f_{a,1} = f_c \pm 1 \text{ MHz.} \\
\text{f}_{a,2} & \quad \text{Second adjacent channel interferer, at least 2 MHz away } (f_{a,2} = f_c \pm 2 \text{ MHz}. \\
\text{f}_{a,3} & \quad \text{Third adjacent channel interferer, 3 or more MHz away } (f_{a,3} = f_c \pm k \text{ MHz, where } k \in [3, 4, 5, \ldots]) \text{. This interference may be up to +40 dB stronger.}
\end{align*}
\]

These four input signals are also shown in figures [S](a-d). Of the \( f_{a,3} \) category, only the most stringent \( (k = 3) \) is shown. In addition to these four interferers, each subplot also depicts the SOI, thermal noise \( (-37 \text{ dB relative to SOI}) \) and the transfer function of the first bandpass filter \( (BPF1) \).

B. Bluetooth: eye diagrams and BER calculation

A Bluetooth signal is modulated using Gaussian Frequency Shift Keying (GFSK) \( [5] \). This effectively
means that a transmitted "1" causes $f_c$ to shift to $f_1 = f_c + f_{\Delta_1}$ and a transmitted "0" results in $f_0 = f_c - f_{\Delta_0}$.

In our Bluetooth modulator $f_{\Delta_0} = f_{\Delta_1}$ and depends on the modulation index $h = 0.28 \ldots 0.35$ (defined in [8]).

After decimation, channel selection filtering and mixing, the output signal $\{h[n']\}$ of the digital channel selection system is processed by a demodulator using an FM to AM conversion function $F_{\text{FM-AM}}(\cdot)$ [5]. The frequency variations are thus converted back to amplitude variations representing the received bits. Figure 4(a) depicts the converted output signal $F_{\text{FM-AM}}(\{h[n']\})$ of a fully processed input signal $z[n]$ with no interferers. Every 10 samples ($= T_{\text{symbol}} = T_{\text{bit}}$) a decision is made whether the transmitted bit was "0" or "1". In figure 4(b) the resulting (soft) bit sequence is shown. By choosing an appropriate trigger moment and plotting each symbol time $T_s$ on top of the next, a so-called eye diagram is obtained (see figure 4(c)). The x-axis thus depicts 10 sample times, and the y-axis is the amplitude of a signal representing this bit. The eye diagram shows the decision moment that was chosen by the demodulator algorithm. The "open-ness" of the eye diagram will be referred to as $\Delta A_{0,1}$, defined as $\Delta A_{0,1} = A_1 - A_0$ (with $A_0, A_1$ as defined in figure 4(c)). The demodulated received bits are written to an output file to determine the BER by comparing them to the transmitted bits.

C. ACI vs ISI trade-off

By introducing bandpass filters to the system (to remove adjacent channel interference and noise), intersymbol interference (ISI) is introduced to the system. Narrow bandpass filters increase ISI, which can be seen in the eye diagrams as narrowing $\Delta A_{0,1}$ (see figure 4(c)). Thus, the probability of correctly detecting whether a "0" or "1" was transmitted is reduced because of narrow bandpass filters. The ef-
fects of *ISI* are also clearly demonstrated in figures 4(a) and (b). The first 5 soft bits (0 ≤ t ≤ 5 · T_s) are "10101" and each have an amplitude of ≈ 0.4. The next 5 bits (5 · T_s ≤ t ≤ 10 · T_s) are "00111" and have significantly larger amplitudes. Thus, if for instance the "10101" amplitude variations become more and more dispersed by choosing a very narrow filter bandwidth, they can eventually cancel each other out. This relationship works two ways. To reduce bit errors caused by *ISI* (*BER_{ISI}*), the filter bandwidth can be increased. This leads to a reduction of channel selectivity and stronger interference from adjacent channels, increasing bit errors caused by *ACI* (*BER_{ACI}*). Thus, by reducing *BER_{ISI},* *BER_{ACI}* is increased. The sensitivities of both contributions and resulting trade-offs will be analyzed in this paper.

III. System analysis

The performance of the system under review depends mainly on sample rate, signal type (real or complex) and the number of filter coefficients used. The performance figure of the system (assessing computational complexity, as discussed in [7]) is minimized by using several methods. These include polyphase implementation of the filters, and length reduction (trade-offs) of filters operating at higher rates and the choice of real signal processing for Bluetooth mode. The resulting channel selection system is divided into two sections, labelled "USB/LSB selection" and "bandpass filter section" (see figure 3).

The USB/LSB selection section includes two antialias lowpass filters (*LPF_1, LPF_2*) of 16 taps each, to be implemented in polyphase. The Hilbert transformer (*Hb*) is also assumed to be of minimal filter length. It requires at least 50 taps to sufficiently attenuate a +40 dB stronger interferer with carrier frequency *f_a,3 = −f_c*. This is due to pass-band ripple constraints discussed in [7]. The largest remaining bottlenecks are *BPF_1* and *BPF_2*, of which *BPF_1* is the most critical. This is of course due to the facts that it has a variable nature and that it cannot be implemented in polyphase (as there is no subsequent decimation). Thus, the USB/LSB selection section will remain unchanged in the following experiments, while the bandpass filter section is analyzed.

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A. Analysis approach

First, the bandpass filter section will be reduced to a simplified study model. Then, a step-by-step approach is chosen to study the *BER* contributions of the input signals and system components. The knowledge gained from this model will then be used to analyze the original model. There are four worst-case input signals (shown in figures 5(a-d)) that must be processed by the receiver and the maximum allowable *BER* of each test is 1 · 10^{-3}.

B. Bandpass filter section analysis

The bandpass filter section will be analyzed as follows: First, the section is removed entirely to find the *BER_{floor}* that is imposed by the remaining receiver components and input signals of figure 6(a). This is done in experiment 1. Then, in experiment 2, one bandpass filter (*BPF*) is introduced to the system (as shown in figure 6(b)).

For each experiment, 4 input signal constellations are used to quantify the *BER* contributions of *ISI, ACI* and *Noise* separately. The inputs are labelled *A − D* and defined as follows:

- *A* = *SOI* with *f_c* = 2.5 MHz
- *B* = Signal *A* + bandlimited white noise (SNR 17 dB)
- *C* = Signal *A* + adjacent channel interferer
- *D* = Signal *C* + bandlimited white noise (SNR 17 dB)

The bandlimited white noise is added in the analog front-end (figure 1), containing a 7th order Butterworth lowpass filter with a cut-off frequency ≈ 10 MHz. In experiments 1 and 2, the (white noise only) signal to noise ratio is chosen to be 17 dB. For the Bluetooth reference tests (as described in sections II-A and IV-C) a more realistic value of 37 dB (thermal noise) is used. The bandpass filter section configuration in combination with the input signals yields tests denoted by test 1A, test 2B etc.

In experiment 2, two filter parameters will be changed: the passband width *B_p* and transition band width *B_t* (as defined in figure 7). The bandpass filter used in this experiment is an equiripple FIR, designed using the Remez Exchange algorithm. The common parameters for the filter are listed in table 1. Note that the filter order follows from pass- and stopband characteristics defined by *B_p* and *B_t*. The ranges are *B_p* ∈ [0.4 . . . 1.1] MHz and *B_t* ∈ [0.2, 0.4, 0.6] MHz. The filter length reduces quickly by increasing *B_t* (as shown in the table).

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1 The filter coefficients must be updated every time the input signal hops to another frequency (refer to [8] for the hopping sequence).

2 This value is chosen to increase the *BER* and thus the reliability of the *BER* calculations.
Fig. 5
TRANSFER FUNCTION OF BPF1 AND ITS INCOMING SIGNAL $d[n']$, FOR 4 SPECIFIED EXPERIMENTS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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<tr>
<td>Type</td>
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<tr>
<td>Design method</td>
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<tr>
<td>Filter order</td>
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TABLE I
Prototype channel selection FIR filter

IV. Experiments

The resulting $BER$ of each experiment is calculated and analyzed to quantify the contributions of $BER_{ISI}$, $BER_{ACI}$ and $BER_{Noise}$ and their dependence on the filter parameters.

A. Experiment 1

As is to be expected, without noise or interference the total $BER$ of test 1A ($BER_{Total,1A}$) is equal to $BER_{Floor} = 0$.

$$BER_{Total,1A} = BER_{Floor,1A} = 0$$  \(1\)

By adding noise to the input, the total bit error rate is increased by $BER_{Noise}$. The relation now becomes:

$$BER_{Total,1B} = BER_{Floor,1A} + BER_{Noise,1B}$$  \(2\)
The result of the test is a $BER_{Total,1B} = BER_{Noise,1B} \approx 0.37$ (remember that $SNR_{Noise} = 17$ dB). To find the influence of adjacent channel interference to the total $BER$ test $1C$ is done. The expected relations are shown in eq. 3.

$$BER_{Total,1C} = BER_{Floor,1A} + BER_{ACI,1C} \quad (3)$$

The result of this test is $BER_{Total,1C} = BER_{ACI,1C} \approx 0.50$. Thus, without a filter the adjacent carrier destroys the ability to correctly receive a bit. The final test $1D$ was formulated to find other factors or correlations influencing $BER_{Total,1}$ that are not in eqs. 1-3. Unfortunately, the disruption of the $BER$ by the adjacent channel prevents this. Clearly, a $BPF$ is required in the system.

B. Experiment 2

Now that a filter is added to the system, $BER_{ACI}$ must be sufficiently reduced. By reducing too much, $BER_{ISI}$ is added to the equation. The relations that are expected to apply in test $2A$ are shown in eq. 4.

$$BER_{Total,2A} = BER_{ISI,2A} \quad (4)$$

The resulting signal has severe $ISI$ for narrow filters, but without added noise or interference the total $BER$ remains 0 for most simulations. This is illustrated in figure 8(a) where $\Delta A_{0,1}$ is nearly zero, but not quite. Thus, $BER_{Total,2B}$ is expected to be a "lifted" version of $BER_{Total,2A}$. Its constituent contributions are shown in eq. 5:

$$BER_{Total,2B} = F_{No} \cdot BER_{Noise,1B} +$$
$$+ F_{ISI} \cdot BER_{ISI,2A} \quad (5)$$

Here, $F_{No}(B_p, B_t)$ and $F_{ISI}(B_p, B_t)$ are assumed to be scaling functions that depend on the filter characteristics. This dependency is also assumed to hold for $F_{ACI}(B_p, B_t)$ in eq. 6. The resulting $BER$ due to $ISI$ and noise is shown in figure 8(b). The contribution of $BER_{ISI}$ is best shown by the $BER_{Total,2B}$ of the sharpest filter with $B_t = 0.2$ MHz. It is clear that such a narrow transition band (which is also responsible for high filter orders) should not be used with $B_p$ values $\leq 0.7$ MHz.

Now, to find the contribution of $ACI$, the noise is removed from the input signal and replaced by an adjacent carrier $f_a$. The total $BER$ is now expected to behave as in eq. 6.

$$BER_{Total,2C} = F_{ISI} \cdot BER_{ISI,2A} +$$
$$+ F_{ACI} \cdot BER_{ACI,1C} \quad (6)$$

From these results, the $BER_{Total,2B}$ values will be subtracted to obtain an estimation of the $ACI$-only contribution. This is shown in figure 8(c). For $B_t = 0.4$ MHz, the adjacent channel interference becomes a problem for $B_p > 0.9$ MHz. The corresponding eye diagram of figure 8(d) clearly shows that when the interferer is allowed to penetrate even further into the passband, $\Delta A_{0,1}$ will be reduced to 0. The final test 2D will be used to find the optimal values for $B_p, B_t$ concerning the $ACI, ISI$ trade-off. Furthermore, discrepancies between the assumed $BER$ contributions and the actual simulation results can be found. The expected relations are given in eq. 7.

$$BER_{Total,2D} = F_{No} \cdot BER_{Noise,1B} +$$
$$+ F_{ACI} \cdot BER_{ACI,1C} +$$
$$+ F_{ISI} \cdot BER_{ISI,2A} \quad (7)$$
(a) ISI-only eye diagram ($B_p = 0.6, B_t = 0.2$)

(b) BER (ISI and Noise)

(c) BER (ACI)

(d) ACI-only eye diagram ($B_p = 0.9, B_t = 0.4$)

(e) Total

(f) Critical points

Fig. 8
EXPERIMENT 2 RESULTS
The optimal passband width is thus where the SOI remains unaffected by the filter for higher frequencies, but restricting the adjacent channel from interfering too much. This can be illustrated by a close-up of the SOI and possible interferers. Disregarding the noise for a moment, figure 8(f) shows the points where adjacent interferences "enter" the SOI (α, β, γ). These points can be seen as the end of the passband, as interference starts there. By choosing a smaller passband, the SOI is attenuated unnecessarily. Bt must then be chosen in such a way that it does not attenuate the outer lobes of the SOI too much, but sufficiently reduces adjacent channel interference (denoted by the points δ, ε, ζ). The results of this trade-off for the current input signals are shown in figure 8(e). The optimal BPF for these input signals is configured using $B_p = 0.7$ MHz and $B_t = 0.4$ MHz. This filter yields the minimal BER. A filter with $B_t = 0.6$ MHz is shown to perform significantly worse, due to excessive adjacent channel interference.

There is a discrepancy between the sum of individual BER contributions and BER_Total of test 2D. Plotting these discrepancies for the entire range shows that they are within 10% for $B_p = 0.6$ to 0.8 and all $B_t$. Thus, strong correlations exist between $F_{ISI}, F_{ACI}$ and $F_{Noise}$ in the regions where either ISI or ACI are dominant. This can be explained by the eye diagrams shown in figures 8(a) and (d). Obviously, when filter characteristics reduce the open-ness of the eye such that the noise amplitude is equal to $\Delta A_t$, its influence is much more severe.

C. Original system

Based on the optimal filter parameters found in the previous section ($B_p = 0.7, B_t = 0.4$ MHz), the optimal bandpass filter response can be derived (as shown in figure 9(a)).

The original system (figure 3) contains two bandpass filters and must resist even stronger interferences (e.g., $f_{a,3}$, section II-A) so the combined stop-band attenuation must be $\approx 40$ dB more. By optimizing for computational load, the filter length of BPF1 is reduced (indirectly reducing ISI). Therefore BPF2 is increased in length to compensate for ACI-induced bit errors. The BPF2 filter type is changed to a Hamming window (FIR) filter [6] to increase stop-band attenuation (by reducing constraints on $B_p$ and $B_t$). The filter responses of BPF1 and BPF2 are shown in figure 9(b). Their combined filter response compared to the optimal filter found is shown in 9(c). The current configuration passes all tests, and its parameters are shown in table II.

V. Conclusions

Experiment 1 has clearly proven the need for channel selection filters in case of adjacent carrier interference and high noise levels. During experiment 2 it became apparent that BER_{ACI} and BER_{ISI} behave as expected and that a clear optimum can be found between the two. The influence of the noise level to the total BER was shown to be extra strong in the regions where ISI and ACI are strong as well. The optimal filter parameters found in the previous section were used as guidelines for the original channel selection system. By trading off filter coefficients and ACI in BPF1 (thus, the constraints on BPF1 were relaxed), BPF2 was used to compensate. To reduce this (extra) ACI contribution, a filter design method with more stopband attenuation was chosen.

References


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<td>$A_s$ [dB]</td>
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Note that $B_p, B_t$ could not be used directly as design parameters for BPF2. It was designed by cut-off frequencies 1.85 and 3.15 MHz.
Fig. 9
FILTER TRANSFER FUNCTIONS