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Research

Digital Algorithms for Suppression of Adjacent Channel Interference in FM-Receivers

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Abstract

In this paper, two new digital algorithms were presented, which allow the demodulation of an FM signal in broadcasting, even if there is one adjacent channel interference with overlapping spectra, which is stronger than the desired signal. The first algorithm is the Cross-Coupled-Baseband DPLL (CC-BB-DPLL), which avoids the problem of changing the synchronization. This is a main problem in proposed CC-PLL systems. Much better compensation results are possible by using the second algorithm called Forward Compensation Structure (FCS), which is presented in this paper for the first time. The FCS is a simplified CC-BB-DPLL with internal Intermediate Frequency (IF)-filtering. The complete signal estimation including an adaptive compensation algorithm was presented. All the filter and estimation structures are optimized using FM model processes generated by the Monte-Carlo Method.

Keywords: Algorithm; Interference; Forward Compensation Structure; Monte-Carlo

Introduction

In [1] a digital approach to FM demodulation using Digital Phase Locked Loop (DPLL) techniques is proposed which can be applied as a digital FM stereo receiver for broadcasting. The algorithms for compensation of adjacent channel interference proposed in this paper also require samples of the quadrature components of the Frequency Modulated (FM) signal, as in equation [1], because this allows the realization at the minimum sampling rate of about $f_A = 1/T_A = 600$ kHz.

$$g(t) = a\,Cos\underbrace{\left[\omega_T t + \phi_a(t)\right]}_{desired\,signal} + b\,Cos\underbrace{\left[\omega_T t + \phi_b(t)\right]}_{adjacent\,channel\,interference} \qquad (1)$$

$$with: \quad \phi_a = 2\pi\Delta F_a \int_t v_a(\tau)d\tau$$

$$\phi_b = 2\pi\Delta F_b \int_t v_b(\tau)d\tau + 2\pi.\underbrace{\Delta f_{TS}}_{carrier\,offset}.t$$

$\Delta F_{a,\,b}$ are the frequency deviations of the two received FM signals. Consider the problem of adjacent channel interference with one interferer having a carrier offset of 100 kHz, which is the 'worst case' in FM broadcasting, because the bandwidth of the two FM signals is about 200 kHz, so there is overlapping spectrum. The amplitudes *a* and *b* of the input signals are constant [3].

In computer simulation the modulated signals $v_a(t)$ and $v_b(t)$ are both stereo-multiplex signals generated by the Monte-Carlo-Method [2].

$$v_{a,b}(t) = \left[u_L(t) + u_R(t)\right] + a_p \overbrace{Sin\omega_p t}^{pilot} + \left[u_L(t) - u_R(t)\right]Sin2\omega_p t \qquad (2)$$

$$with: \quad u_{L(R)}(t) \cong left\ (right)\ channel$$

The effective frequency deviation ΔF_{eff} corresponds to the rms of $v_{a(b)}(t)$ without pilot and is chosen identical for both input signals. The sampling rate is f_A = 600 kHz. The influence of the analogue-to-digital conversion at the input and finite-word-length effects is not considered.

1. The CC-BB-DPLL

In [3], [4], [5] a novel FM-demodulator called CC-PLL is proposed, which has the capability to suppress adjacent channel interferer. Using Baseband-DPLL (BB-DPLL) [1] for the phase estimation results in the digital Cross-Coupled-Baseband – DPLL (CC-BB-DPLL) in Figure 1, which illustrates the principle of a Cross-Coupled demodulator. It consists of two separate FM-demodulators including phase- and amplitude-estimation. They are inter-connected in order to permit the demodulation of the stronger and the weaker of both input signals.

One main problem in a Cross-Coupled demodulator is its symmetry; the BB-DPLL in "structure A" can either lock onto the adjacent channel interferer or onto the desired signal. Also the synchronization can change, if the instantaneous frequencies of the two received signals cross. The phase estimation shown in Fig. 1b avoids this effect by using a non-linear limiter NL at the output of the phase error estimation Gφ (z), which is a simple low pass filter of order two. The BB-DPLL is of order one with $F(z) = M = 1.0$. So the input signal of Gφ is the demodulated signal which is a first estimation of the instantaneous frequency of the desired signal. Using a low pass filter $G\varphi(z)$ with the property $G\varphi\ (\Omega = 0) = 1$, a static phase error between ϕ_a and ϕ_a^ is avoided, if there is a frequency offset. This is of importance in structure B. The *N L* is just a limiter with $\Delta\Omega_{TS}/2$ as an upper bound in structure A and a lower bound in structure B, if the carrier-offset Δf_{TS} is positive. If Δf_{TS} is negative, $(-\Delta\Omega_{TS}/2)$ is the lower bound in structure A and the upper bound in structure B. So an estimation of the sign of Δf_{TS} is necessary.

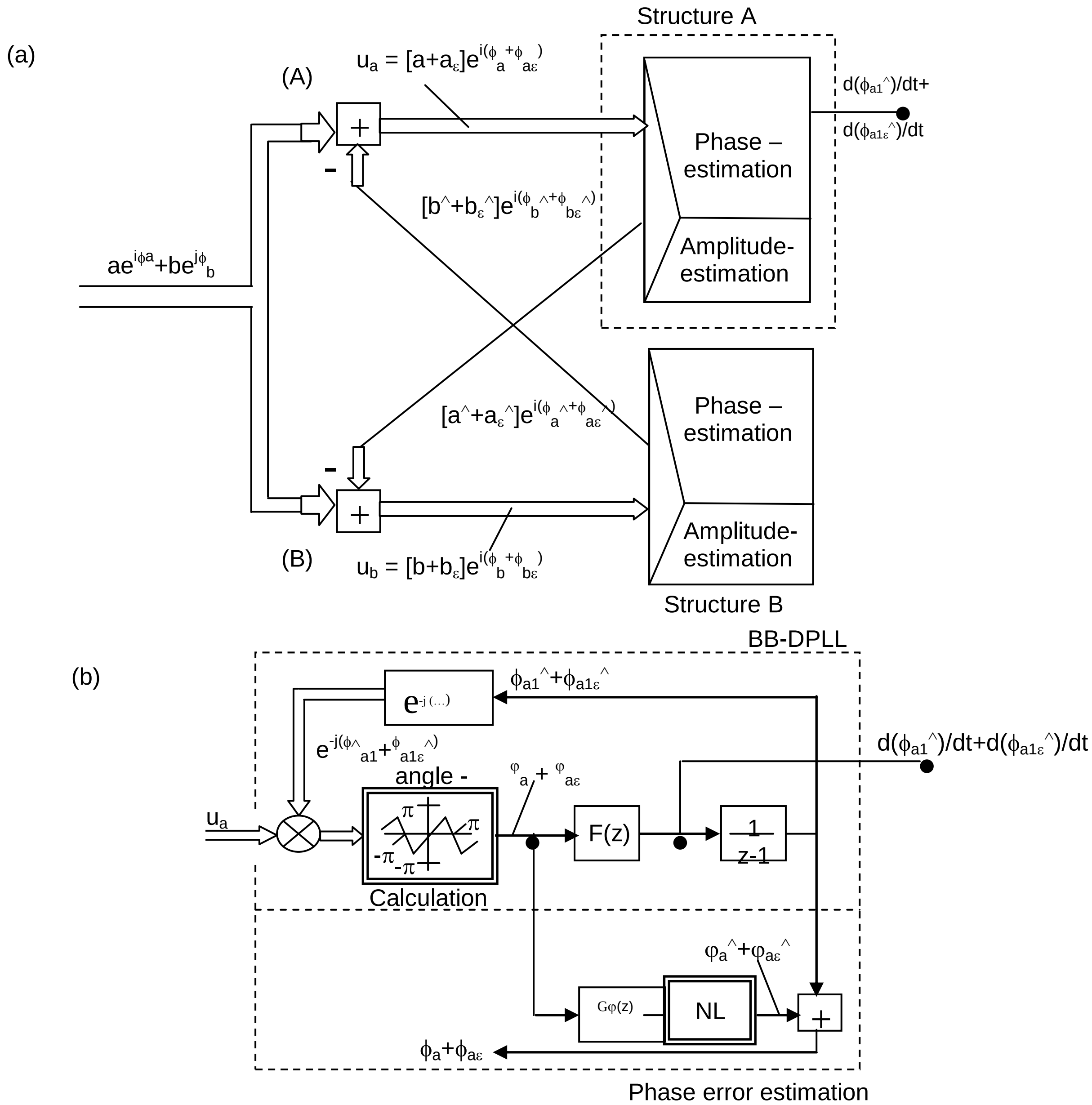

Fig. 1: (a) The CC-BB-DPLL (b) Phase estimation in the CC-BB-DPLL.

Another problem using a CC-BB-DPLL is its feedback structure. In order to get a good compensation result, any delay time should be avoided. This leads to a very large noise bandwidth Ω_R of the phase estimation:

$$\Omega_R = \frac{1}{\pi[G_\varphi(\Omega = 0)]} \int_0^\pi [G_\varphi(\Omega)]^2 \, d\Omega \qquad (3)$$

With the z – transform:

$G_\varphi(z) = \phi\hat{}\,(z) / \phi\,(z) \quad = \quad G(z) + G_\varphi\hat{}\,(z)\,[1 - G)\,(z)]$

$G(z) = \phi_1\hat{}\,(z) / \phi\,(z) \quad = \quad F(z) / [z - 1 + F(z)]$

Figure 2 shows some simulation results. The amplitude estimation is ideal, that is the amount of a and b are known and they are used as estimation results.

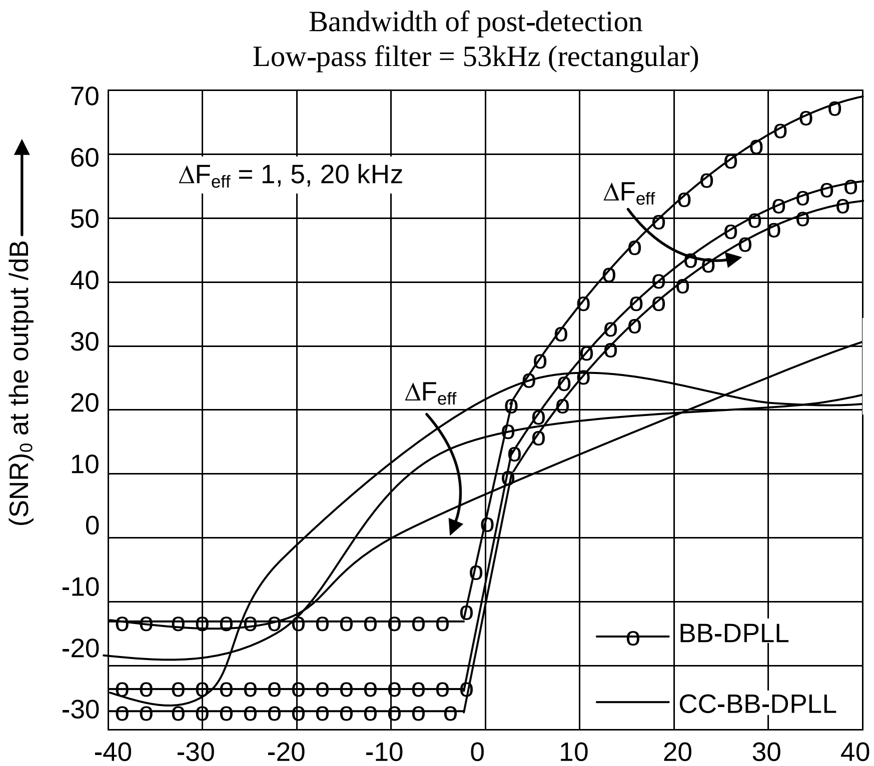

Fig. 2: Output SNR vs. $(SNR)_{HF} = \left(\frac{a}{b}\right)^2 = \rho^2$

The low pass filter $G_\varphi(z) = (b_1 z + b_o)/(z^2 + a_1 z + a_o)$ with $b_1 = 0.3377$, $b_o = 0.3682$, $a_1 = 0.4459$, $a_o = 0.1518$ leads to a noise bandwidth $\Omega_R = 2.0$. As expected in the case of the BB-DPLL, the $(SNR)_o$ is proportional to the $(SNR)_{HF} = (a/b)^2 = \rho^2$. The CC-BB-DPLL has an advantage, when the $(SNR)_{HF}$ is negative, because the compensation still allows the demodulation of the desired signal. But for $(SNR)_{HF} > 0$ dB, we better use the simple BB-DPLL.

2. The FCS

In order to get better compensation results for $(SNR)_{HP} < 0$ dB, the Forward Compensation Structure (FCS) in Figure 3 is introduced. The FCS is simply the CC-BB-DPLL compensating only the adjacent channel interferer. Because no feedback takes place, a delay time in the estimation can now be balanced. The phase estimation includes a BB-DPLL, a filter $H_E(z) = 1/[zG(z)]$ in order to equalize the closed loop transfer function of the phase locked loop, a low-pass filter H_{TP} for suppressing adjacent channel interference, an accumulator, and a phase modulator.

In order to avoid a phase offset between the output signal of the accumulator ($\phi_b + \phi_{b\varepsilon}$) and the input phase ($\phi_b\hat{} + \phi_{b\varepsilon}\hat{}$), the accumulator is part of a second first order BB-DPLL, which is a narrowband-DPLL with $M = 2^{-4}$ (Implicitly embeded in Fig. 3). This DPLL is detuned by ($\phi_b\hat{} + \phi_{b\varepsilon}\hat{}$). Its input signal is also $(b+b_\varepsilon)$ exp [j ($\phi_b\hat{} + \phi_{b\varepsilon}\hat{}$)] taking into account the delay time of the first BB-DPLL, H_E and H_{TP}.

In order to understand the principle of the FCS, let us reasonably suppose that there is an increasing amplitude ratio b/a beginning at $b/a = 0$. First no compensation takes place, because structure B locks onto the desired signal. When the ratio is $(b/a) \approx 1$, the 1F-filter with the frequency response shown in Fig. 4 still allows the demodulation of the desired signal until the BB-DPLL in structure B locks onto the adjacent channel interferer. Then compensation takes place and demodulation of the desired signal is still possible, even for $(SNR)_{HF} < 0$ dB.

The algorithm used for adaptive compensation first estimates the mean value of the output signal d($\phi_b\hat{}$)/dt+ d($\phi_{b\varepsilon}\hat{}$)/dt of structure B by a *RC* low pass filter with $f_{3dB} = 2$ kHz. Compensation is performed, if the absolute mean value is greater than $\Delta\Omega_{TS}/2 = \pi / 6$ ($\Delta f_{TS} = 100$ kHz; $f_A = 600$ kHz) during the sequence of forty samples. This ensures, that the BB-DPLL in structure B has locked onto the adjacent channel interferer. Compensation is switched off, if this condition is not fulfilled; thus the switch off happens very fast.

The BB-DPLL in structure B is of order two with an active loop filter $F(z) = M(1 + x/(z-1))$ ($M = 0.67$; $x = 0.1675$). The BB-DPLL in structure A is of order one with $F(z) = M = 0.5$. The low-pass filter H_{TP} shown in fig. 5 is a special one. The low—pass filter H_{TP2} is a cascade consisting of an all-pass filter with poles at $z_{001} = 0.7340$; $z_{002,3} = 0.7866 \pm j\ 0.1531$; $z_{004,5} = 0.7494 \pm j\ 0.3709$ and a Chebyshev low-pass filter of order 6 with equal ripple of 21.2 dB in the stop band. Its cut-off frequency is $f_g = 70$ kHz. If K_2 is one, H_{TP} corresponds to H_{TP2}. If K_2 is zero, no low pass filter is used. In order to get the best $(SNR)_0$ at the output of structure A, K_2 is chosen as a function of the amplitude ratio.

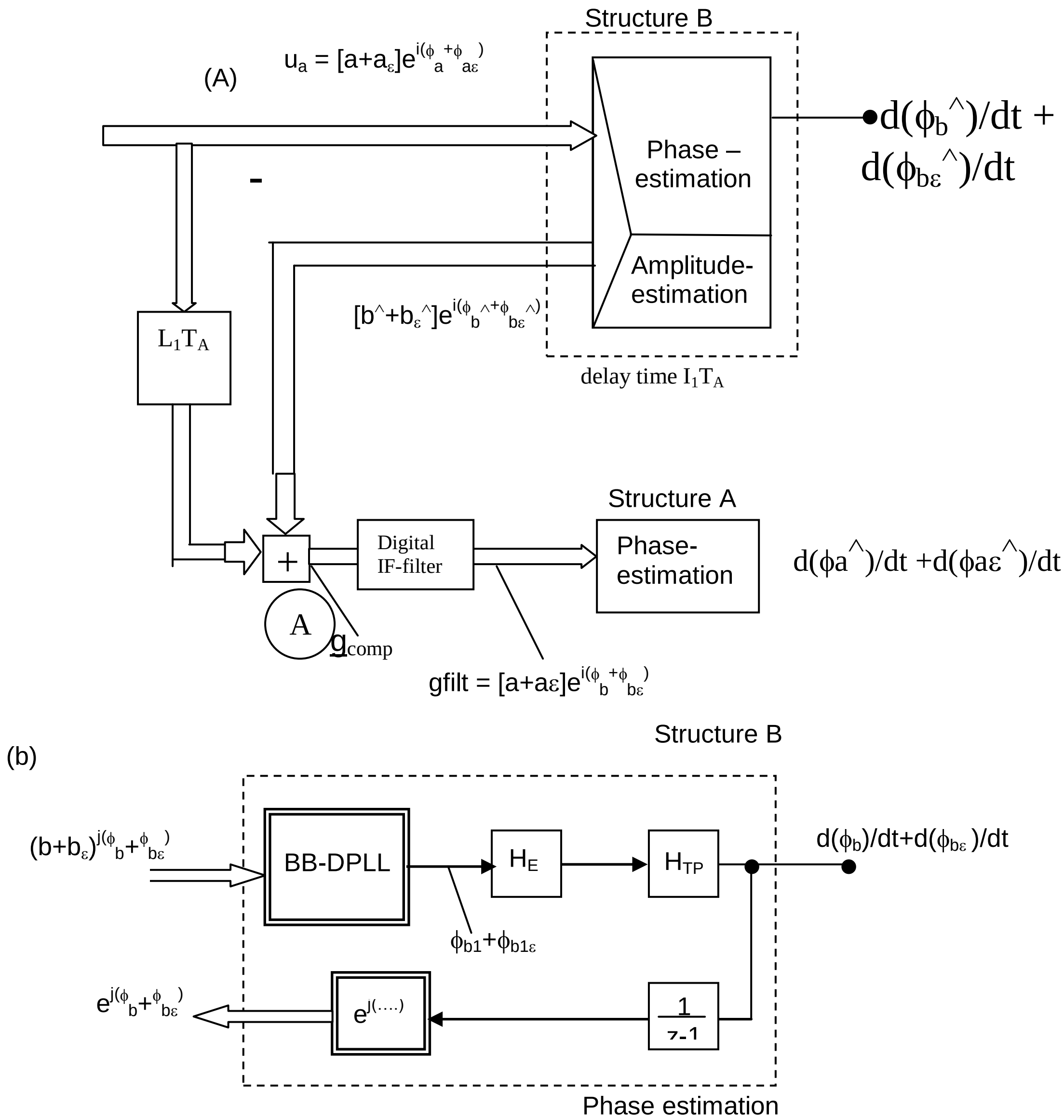

Fig. 3: (a) The FCS; (b) Phase Estimation in the FCS

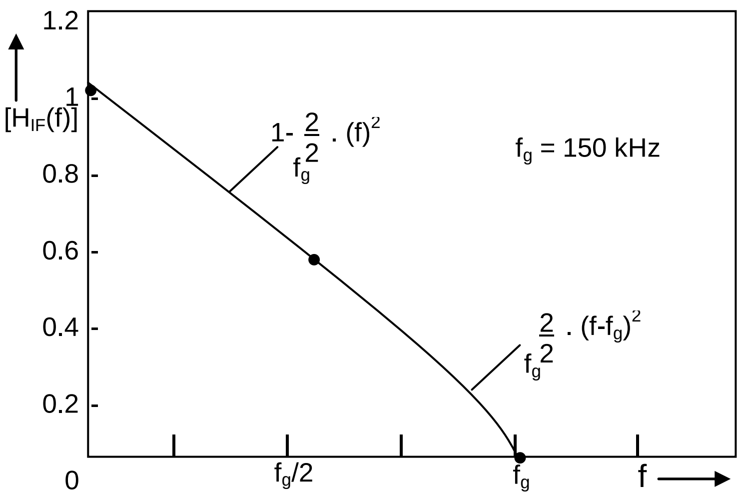

Fig. 4: Frequency response of the IF-filter

$$K2 = \begin{cases} \alpha_o\left[\rho^2 - 0.01\right] & if \quad 0.1 \le \rho \le 1 \\ \alpha_o & if \quad \rho > 1 \\ 0 & if \quad \rho < 0.1 \end{cases} \tag{4}$$

with $\alpha_0 = 0.84$.

So if the $(SNR)_{HF}$ is less than -20 dB, it is better not to use the low pass filter H_{TP}, because noise shaping at the input of the IF-filter takes place.

In order to understand the noise shaping, assume an ideal amplitude estimation. Therefore,

$$g_{comp} = ae^{j\phi_a} + be^{j\phi_b} - be^{j(\phi_b - \phi_{b\varepsilon})} \tag{5}$$

If no low pass filter H_{TP2} is used ($K_2 = 0$), then

$$\phi_b{}^{\wedge} + \phi_{b\varepsilon}{}^{\wedge} = \phi_b + \phi_{b\varepsilon} \tag{6}$$

which leads to

$$g_{comp} = ae^{j\phi_a} + be^{j\phi_b} - be^{j(\phi_b - \phi_{b\varepsilon})} \tag{7}$$

Making the substitution

$$\begin{aligned} ae^{j\phi_a} + be^{j\phi_b} &= be^{j\phi_b}(1 + qe^{j(\phi_a - \phi_b)}) \\ &= re^{j\phi_{b\varepsilon}} \end{aligned} \tag{8}$$

equation (7) becomes

$$g_{comp} = a\left(\frac{r-1}{r}\right)e^{j\phi_a} \quad + \quad b\left(\frac{r-1}{r}\right)e^{j\phi_b} \tag{9}$$

If $\rho << 1$, then

$$\frac{r-1}{r} \approx \rho Cos(\phi_a - \phi_b) \tag{10}$$

Inserting (10) in (9) yields:

$$\begin{aligned} g_{comp} &\approx \left(\frac{a}{2}\right)e^{j\phi_a} \quad + \quad \left(\frac{b}{2}\right)e^{j(2\phi_b - \phi_a)} + a\rho Cos(\phi_b - \phi_a)e^{j\phi_a} \\ &\approx \left(\frac{a}{2}\right)e^{j\phi_a} \quad + \quad \left(\frac{b}{2}\right)e^{j(2\phi_b - \phi_a)} \end{aligned} \tag{11}$$

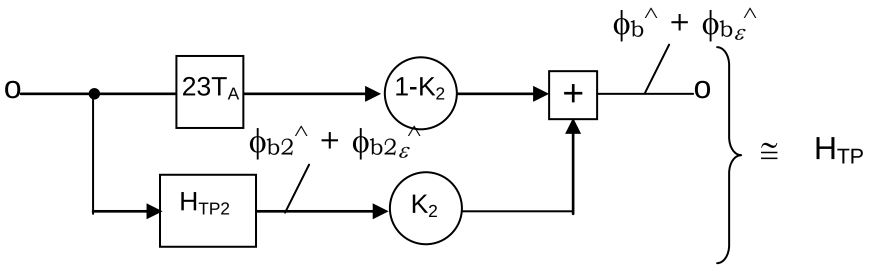

Fig. 5: Low pass filter H_{TP}

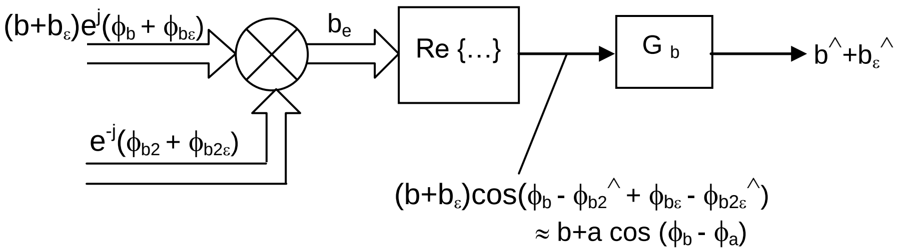

Fig. 6: Amplitude estimation

The result is an adjacent channel interferer at twice the carrier offset $2\Delta f_{TS}$ with the same amplitude as the desired signal. This modified interferer will be nearly suppressed by the digital IF – filter.

The last topic for treatment is the amplitude estimation in Fig. 6. Using the phase estimate $\phi_{b2}^{\wedge} + \phi_{b2\varepsilon}^{\wedge}$, b_ε will be a narrow-band FM signal. So the real part of b_ε is nearly $(b+b_s)$. In order to suppress the influence of the adjacent channel interference in the amplitude estimation, $\phi_{b2}^{\wedge} + \phi_{b2\varepsilon}^{\wedge}$ is generated by accumulating the output signal of H_{TP2} in Fig. 5 in a third BB-DPLL of order one with $M = 2^{-4}$ in the same manner as described above for the phase estimate ($\phi_{b2}^{\wedge} + \phi_{b2\varepsilon}^{\wedge}$). G_b is a Chebyshev low-pass filter of order 3 with equal ripple of 40 dB in the stop band. The cut-off frequency is 30 kHz. The delay time of G_b is balanced at f = 1 kHz, in order to get good compensation results for low frequency amplitude variations. The output signal of the amplitude estimation is applied to the adjustment of the low pass H_{TP} as described in equation (4).

Squaring the input signal,

$$[ae^{j\phi_a} + be^{j\phi_b}]^2 = a^2 + b^2 + 2abCos(\phi_b - \phi_a) \tag{12}$$

If we assume $b + b_\varepsilon \approx b$ and we first subtract $(b^{\wedge} + b_\varepsilon^{\wedge})^2$ and then divide by $(b^{\wedge} + b_\varepsilon^{\wedge})^2$ in Eq. [12], the result is

$$y \approx \rho 2 + 2\rho Cos(\phi_b - \phi_a) \tag{13}$$

Using a RC low pass filter with f_{3dB} = 2 kHz for the suppression of the second term in equation (13), a good estimate of the desired ratio q^2 in equation (4) is possible. Figure 7 shows some simulation results. The input signals are the same as used in Figure 2 in the case of ΔF_{eff} = 20 kHz.

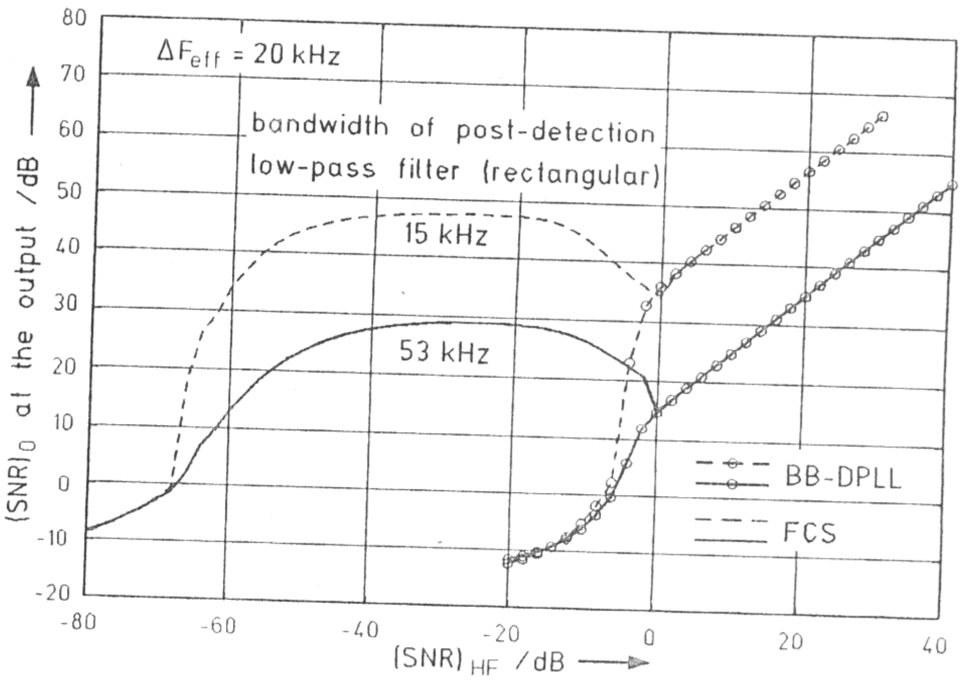

Fig. 7: Output SNR vs. $(SNR)_{HF} = \left(\frac{a}{b}\right)^2 = \rho^2$

3. Conclusion

Two new algorithms are presented which allow the demodulation of the weaker of two FM signals with overlapping spectra using compensation methods. The starting point of the research was the CC- PLL introduced in [3], [4] and [5]. The CC-BB-DPLL and the FCS avoid changing the synchronization which is a main problem in the CC-PLL. Furthermore, because any delay time can be balanced a suitable design can be found, if we use the FCS.

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