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Improved π -Delayed Harmonic Rejection N-Path Mixer for Low Power Consumption and Multistandard Receiver

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Abstract— The purpose of this paper is to study the complexity of the Harmonic Rejection N-Path Mixer HRNPM based receivers. In this framework, a HR-NPM architecture has been discussed based on the π -delayed driving signals. The harmonic rejection and the system complexity have been improved with this new architecture such that the Harmonic rejection achieved by this structure is the same of the conventional HR-2NPM and the system complexity has been reduced by half, where as the number of the gain stages and the switches has been reduced. In this context, a 5-path mixer is thus proposed. It allows rejecting up to the 8th harmonic with only 2 differential gain stages by appropriately implementing the switches whereas only the 6th harmonic would be rejected with conventional HR-8PM topologies having 3 amplifier stages. Our structure shows high resilience to clocks overlapping effects.

Keywords— N-Path mixer, mixer-first, receiver, software-defined radio, harmonic rejection, blocker, Overlapping.

I. INTRODUCTION

In the last decade, the internet of things (IoT) has gained an incredibly high interest due to the large range of opportunities it can offers in terms of applications. Accordingly, to the application requirements, standards might be chosen appropriately. As an illustration, Table I summarizes the allocated frequency bands for the most usual standards, showing some needs for reconfigurability ranging between 0.38 and 2.48 GHz, (0.36 and 2.6 GHz by considering image signal effect).

In parallel, the software defined radio (SDR) has drawn attention because of its versatile hardware, which adapts its reception and transmission profiles by software [1]. One of the most important features of SDR being its tunability, the receiver components must also be able to be tuned. Fig.1 shows the block diagram of a typical multi-standards band-pass filter (BPF) first receiver. For each standard, a specific filter, usually consisting in lumped LC components has to be designed. A solution to overcome this multitude of receiving paths consists in having the mixer to play both mixing and filtering role. A single component, N-Path mixer, fulfills this complex requirements with two possible architectures, LNA-first [2] or mixer-first [3], switched through accurate clock LO

TABLE I. FREQUENCY BANDS ADDRESSED BY VARIOUS STANDARDS IN THE 300MHZ-2.5GHZ RANGE

Standard	Owner	Frequency (MHz)
Zigbee	Zigbee Alliance	868-868.6 (Europe) 902-928 (US) 2400-2483.5 (Global)
T-GSM	GSM	330 410 810 900
Z-Wave	Sigma Designs	868.42 (Europe) 908.42 (US)
LTE-M	GSM-A Carriers	450-2350 (uplink)
Lo-RaWan	LoRa Alliance	433, 868 (Europe) 915 (US) 470 (China)
IEEE 802.11af	Open-IEEE Certified	470-710 (Digital Dividend)
IEEE 802.11ah	Open-IEEE Certified	850 (Europe) 900 (US) 700 (China)
IEEE 802.15.4	Open-IEEE Certified	868.3 (Europe) 902 (US) 2400 (Global)

signals. In the conventional NPMs, unfortunately, some harmonics are introduced in the IF response due to LO sampling, thus degrading the width of the rejection band. Harmonic Rejection N-Path Mixer (HR-NPM), based on LNA-first architectures, are an excellent solution in the framework of SDR with multi-standards receivers [2].

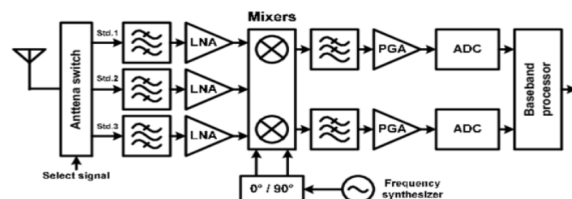


Fig. 1 Block diagram of a typical multi-standards receiver architecture

Fig. 2 illustrates the combination of wide-band LNAs together with the appropriate LO clock signals of frequency F_{LO} in the generic topology of HR-NPMs. It can be demonstrated that rejection by conventional HR-NPMs can theoretically cover a band of interest from 1 to $(N-1) \cdot F_{LO}$ [4]. Anyway, efforts have to be committed on reducing the complexity of HR-NPMs while keeping the highest rejection band as possible, with a strong resilience to LNAs gain and phase mismatches as well as a strong resilience to clocks overlapping.

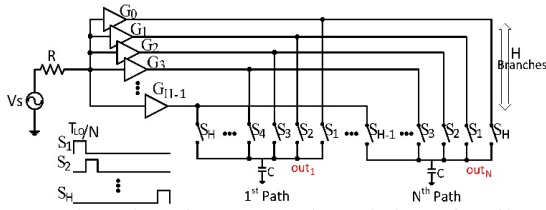


Fig. 2 Generic topology of a HR-NPM, where H is the number of branches and N the number of paths.

In this context, after briefly reminding the principle of the HR-NPMs and the notion of Effective Local Oscillator (EFLO) in section II, this paper describes two low complexity and high harmonic rejection band receivers that enable to cover the multi-standards band from 0.36 to 2.6 GHz, based on novel N -path mixers architectures, in section III. The first one has already been introduced in [5], showing great interest in terms of complexity and HRR as compared to the conventional HR-8PM while keeping a similar resilience facing LNA outputs mismatching. Meanwhile the second one is a new architecture that presents even lower complexity. Harmonic rejection performance of both structures is studied, on the basis of the signal properties, by considering the overlapping clocks effects. Section IV provides a discussion before concluding.

II. CONVENTIONAL HARMONIC REJECTION N-PATH MIXERS

A. HR-NPMs General Architecture

To get a well point of view about the HR-NPM architectures, a general architecture has been shown in Fig. 2, the HR-NPM architecture composed of N Paths, each path consists of H branches symbolized by switches and connected by a single loaded BB capacitor, where as each branch is the following to LNA with the different gain level. The procedure to get the gain coefficients has been studied before with different criteria in [4]. In this work, the gain coefficients are the result of samples of sine wave with N samples, then the effective LO of each path is equivalent to a sine wave sampled by N samples. These coefficients are given by the following series defined for $0 \leq h \leq H-1$:

$$G_h = \sqrt{2} \sin\left(\frac{2\pi}{N}\left(1+h\frac{N}{H}\right)\right) \quad (1)$$

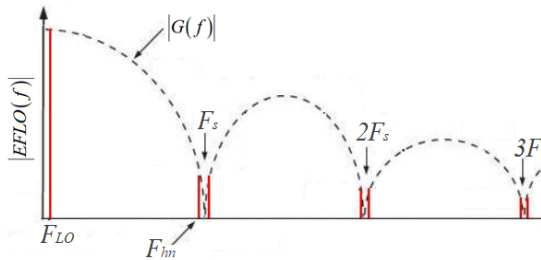


Fig. 3 Frequency response of the effective LO, $eflo(t)$. $F_{hm} = nH.F_{LO} \pm F_{LO}$

The output signal ($out_n(t)$) spectrum at each path is given by (2):

$$OUT_n(f) = V_S(f) * EFLO(f) \quad (2)$$

with:

$$|EFLO(f)| = \frac{1}{2} H.F_{LO}.T_s \text{sinc}(\pi T_s f) \sum_{n=0}^{+\infty} \delta(f - nH.F_{LO} \pm F_{LO}) \quad (3)$$

The first harmonics in the $eflo(t)$ response of the general HR-NPM architectures occurs at $F_{hm} = n.F_s \pm F_{LO} = (N \pm 1).F_{LO}$ as shown in Fig. 3, these harmonics also known by folded harmonics which having hard impact on the fundamental one, these folded harmonics must be rejected before the mixing. Based on this figure, the HR band directly affected by the paths number N which can be improved by increasing N .

B. HR-8PMs Architecture

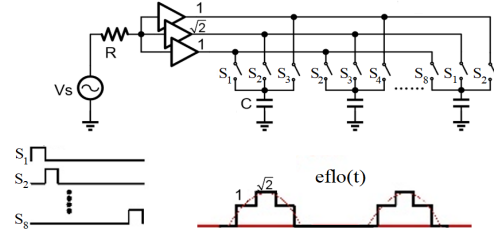


Fig. 4 Structure of a HR-8PM LNA-first receiver with a Single structure.

The principle of the HR-8PM structure is to generate an effective LO $eflo(t)$ near as possible to the pure sine wave with 8 samples levels, where as these samples came from sampling the sine wave signal by 8 samples, this induces 3 positive gain levels with gain ratio $[1; \sqrt{2}; 1]$, the schematic of this topology (single structure) is composed from 8 paths where each one consists of 3 switches formed by simple transistors controlled from the gate porte and followed by 3 gain stages, respectively, as shown in Fig. 4 [1] and [2]. Based on the property of Fig. 3, the 1st harmonic in the effective LO response of this architecture occurs at the 7th order ($8-1$), thanks to this structure for eliminating the 3rd and 5th harmonics orders from its $eflo(t)$ response EFLO.

C. Complexity Analysis

Fig. 5 shows the relation between the HR band with the overall complexity architecture needed to achieve that band. On the basis of [4], the HR-8PM achieves HR band until 6th harmonic while requiring 3 LNAs and 48 switches. The HR-10PM increased the HR band till the 8th harmonic while the number of LNAs and switches of this structure increased to 4 and 80, respectively. The 9th harmonic has been eliminated with HR-12PM architecture to improve HR band until the 10th harmonic order, while the LNAs and switches number have been increased to 5 and 120, respectively. The main challenge appears by increasing HR band while decreasing the overall complexity system.

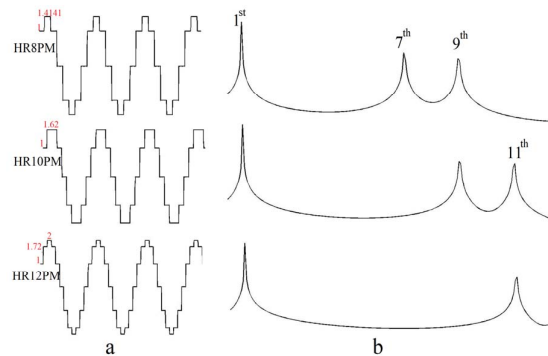


Fig. 5 Effective LO for HR-NPMs with different $N=8, 10$ and 12 in (a) Time domain. (b) Frequency domain.

III. PROPOSED HARMONIC REJECTION-N-PATH MIXERS

To cover the band 0.36-2.6 GHz, an HR-NPM with $N > 8$ ($2.6/0.36 > 7$) has to be implemented. HR-10PM rejects the 7th LO harmonic order with, as a drawback, a large complexity. A novel architecture with lower complexity was proposed in [4] to challenge these features.

A. HR-5PM with equivalent HR-10PM Effective LO

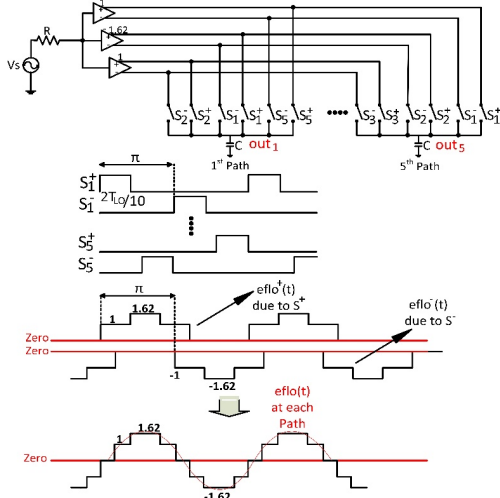


Fig. 6 π -delayed architecture with $H = 10$ and $N = 5$. $H_g = 3$. $S_w = 30$.

1) Architecture

The differential structure with $H=2N$; where $N=(2i+1)$ is the number of paths; is a solution to increase the HR avoiding complexity and maintaining a low VCO frequency. The EFLO signal with $2N$ samples is synthesized due to the combination of the N -paths, switches and their driving signals. In order to control these switches, we need two driving signals Sh^+ and Sh^- for positive and negative branches respectively. Sh^+ generates a positive effective LO $eflo^+(t)$, while Sh^- generates a negative one $eflo^-(t)$ (Fig. 6). Where as the delay between these two driving signals is π . There is a similarity between the EFLO signal at each paths and the standard EFLO of a conventional HR-NPM where $H=N=2.(2i+1)$ after the appropriate gains G_h combination. A remarkable advantage for clock generation is that the VCO frequency of the π -delayed architecture is only $N.FLO$ instead of $2.N.FLO$. Also the number of differential gain stages, H_g , is $H_g = (H_{ef}/2 - 1)$ where H_{ef} is the number of non-null coefficient and $H_{ef} = H - 2$, and the total number of switches is only $S_w = 2.H_g.N$.

2) Overlapping Clocks Effect

Due to the charge time constants $\tau = R.C$ and the technology limitations, the rising, T_r , and falling, T_f , times of the control clocks are not null and they must be taken into account. Considering rising and falling times leads to appear a phenomena known by overlapping[5].

Fig.7 compares the theoretical effective LO of the proposed structure where $T_r = T_f = 0$ and the same curve when considering the effects of rising-falling clocks time. The percentage of overlapping has been exaggerated for visual effects on Fig. 7.

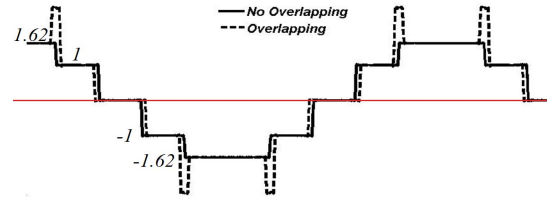


Fig. 7 Effective LO forms with and without overlapping effect of the π -delayed architecture with $H = 10$ and $N = 5$.

The response of the effective LOs with two different overlapping times (supposing that rising and falling time are the same) have been displayed on Fig. 8. As observed, in ideal case, the 1st harmonic occurs at the 9th order while increasing the overlapping affects the 5th harmonic order rejection. As a consequence, even if the idea sounds great, this result avoids to implement this architecture. Intuitively, as the effective LO has to look like a pure sine wave as much as possible, there is general drawback with the π -delayed structure because of the concomitance at the top of the curve of two identical samples. As overlapping is unavoidable in real cases, a smart structure should enable a similar overlapping at each sample edge in order to reduce its overall impact on harmonic generation. The new proposed architecture introduced in the next section overcomes this issue.

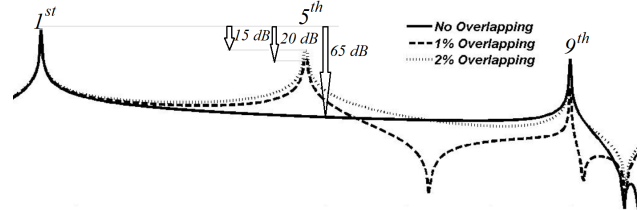


Fig. 8 Response with 3 different overlapping times of the π -delayed architecture with $H = 10$ and $N = 5$.

B. HR-5PM with Non-Return-to-Zero Effective LO

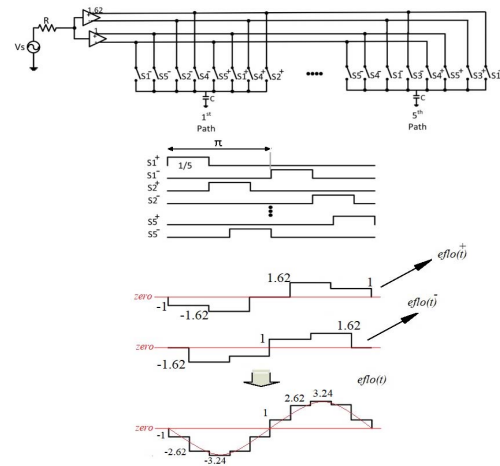


Fig. 9 π -delayed architecture with non-return-to-zero effective LO with $H = 10$ and $N = 5$. $H_g = 2$. $S_w = 40$.

1) Architecture

The way of generating a non-return-to-zero effective LO is also based on the π -delayed clock signals S_h^+ and S_h^- as shown on Fig. 9. The idea is illustrated with five paths but could be extended as well with any HR-NPM as long as $H = 2N$ and $N = (2i+1)$ is an odd number of paths. The $eflo(t)^+$

is exactly the sampled sine wave signal with 5 samples [4],[Table. 1], the response EFLO⁺ eliminates the 3rd and 5th harmonics, while the 1st harmonics occur at the 4th and 6th orders as illustrated in Fig.3 which are even harmonics and can be eliminated by the differential structure, for this reason, the $eflo(t)^+$ has been generated, which is the same of the $eflo(t)^+$ with π phase shift, such that the combination of the $eflo(t)^+$ and $eflo(t)^-$ generates a differential $eflo(t)$ in which the 1st harmonic in its response arises to the 9th order.

Each path is now served through a network of eight switches so that each LNA sees two capacitors in the same time, and this whatever the time. With the preceding approach, because of the switches arrangement, this opportunity of having only two gain stages was not possible, even if two gains only were necessary as well, i.e. 1 and 1.62, to shape $eflo^+$ and $eflo^-$. As a matter of fact, the gain stage of 1 would have seen either one, either two capacitors, depending on the sampling time, which would have been a non-sense in terms of LNA design specifications, so that three gain stages were necessary. Thanks to the new approach, the complexity is now decreased in terms of the number of gain stages, at the very little expense of a larger number of switches, which does not account for power consumption. As a general way, the number of differential gain stages, H_g , is equal to only half the number of paths, N , reduced by one, $H_g = (N-1)/2$. The total number of switches is now $S_w = 4.H_g.N$.

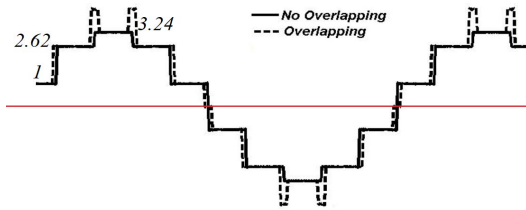


Fig. 10 Non-return-to-zero effective LO forms with and without overlapping effect of the π -delayed architecture with $H=10$ and $N=5$.

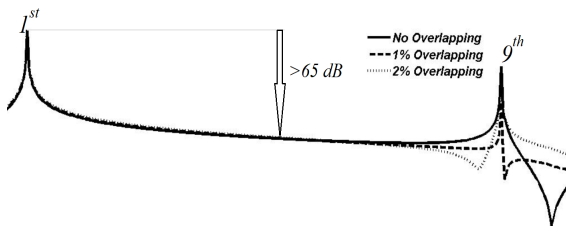


Fig. 11 Response with 3 different overlapping times of the π -delayed architecture with non-return-to-zero effective LO with $H=10$

2) Overlapping clock effects

As a consequence, the form of the effective LO will be changed with respect to the overlapping while the HRRs will not be affected. Fig. 10 shows the ideal and non-ideal with overlapping effective LO forms. It can be observed now a much more regular effect of the rising and falling time on the overlapped curve. The response to two different overlapping (1% and 2%), demonstrates in Fig. 11 that the HRR for the 3rd, 5th and 7th will be kept the same. This is another advantage of the novel implementation of switches since in the real case the overlapping phenomenon cannot be avoided.

IV. DISCUSSION

For $N=5$ and $H=10$, the presented solution involving a non-return-to-zero effective local oscillator needs to implement $H_g=2$ differential LNAs and $S_w=40$ switches whereas the equivalent $H=N=10$ standard structure requires $H_g=5$ and $S_w=100$ for the same total harmonic rejection band. In addition, the performance of the π -delayed structure is theoretically compared to a standard HR-8PM (Table II) corresponding to $H=N=8$, in terms of complexity, rejection band, resilience to clocks overlapping, and resilience to LNA outputs phase and gain mismatches (see [4]).

TABLE II. PERFORMANCE COMPARISON BETWEEN THEORETICAL HR-8PM AND THE NON-RETURN-TO-ZERO, π -DELAYED HR-5PM

	Conventional HR-8PM	Proposed HR-5PM of III.B
LNAs	3	2
Switches	48	40
Rejected harmonics	3 rd , 5 th	3 rd , 5 th , 7 th
Resilience LNA mismatch	Similar (see [4])	
Resilience to overlapping	Similar [this work]	

V. CONCLUSION

This paper presents a theoretical useful study concerning the design of conventional N-path mixers based on LNA-first receivers. The odd numbers of paths justify the interest in the symmetrical properties in terms of sampling gain coefficients, where this symmetry is used in the π -delayed architecture to increase the harmonic rejection band avoiding increasing in the system complexity and maintaining a low VCO frequency. The solution is a 5-path mixer with a specific switches implementation, only two differential amplifiers are enough to obtain a high harmonic rejection up to the 8th harmonic, with the same flexibility as HR-8PM regarding the clocks overlapping and LNA mismatching.

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