# Towards Suppression of All Harmonics in a PolyPhase Multipath Transmitter

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*Abstract*— **This work proposes a direct conversion transmitter architecture intended for cognitive radio applications. The architecture is based on the poly-phase multipath technique, which has been shown to cancel out many of the harmonics, sidebands and nonlinearity contributions of a power up-converter using a large number of signal paths. This work proposes a design only using 8 paths which is able to achieve <-40dBc harmonic suppression for all harmonics, including the dominant 7th and 9th harmonic. This is done with a combination of duty cycle control of the Local Oscillator (LO) waveform and tunable filtering with only a first order roll-off.**

***Keywords; Harmonic Rejection Mixer, Distortion Cancellation, Power Upconverter, Cognitive Radio, Software Defined Radio***

## I. INTRODUCTION

Regulators in North America and Europe are opening up the RF spectrum for devices that can operate where-ever there is free spectrum available in a certain RF band. In order to meet regulatory requirements it is crucial that these devices do not interfere with incumbent users of the frequency spectrum. As shown in [1], one of the challenges for a portable cognitive radio transmitter is that the out of band emissions (OOB) beyond the adjacent channel have to be less than 53dB relative to the desired signal power of 20dBm. In terms of absolute power this means that the OOB emissions should be ≤ -33dBm (5MHz signal bandwidth assumed as in [1]). For an output power of 0-10dBm, which can be enough for a portable device, the OOB emissions should then be 33dB-43dB respectively below the desired signal. In frequency agile transmitters, achieving high OOB suppression is a challenge since tunable high-Q RF-band-pass filters are difficult to implement, while switching mixers produce many strong harmonics. Assuming a square-wave Local Oscillator (LO) with ideal edges and 50% duty cycle, only the 100th harmonic of the LO is at a level of -40dBc.

Techniques such as Harmonic Rejection Mixing [2] have been shown to suppress the 3rd and 5th harmonic of the LO but the 7th and higher harmonics are still there. If Cognitive Radio Transmitters have to cover multi-decades of frequency band [3], [4], then higher harmonics need to be suppressed. The multipath architecture [5] is attractive as it has been shown to cancel both a large number of LO harmonics and also distortion products from the power up-converter (PU). The challenge is to maintain or improve the harmonic

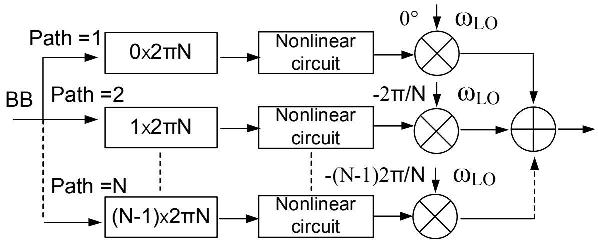



Figure 1. Basic Principle of Polyphase Multipath Up-conversion

suppression level while minimizing complexity.This paper proposes a multipath architecture which relies on just 8 paths to suppress all the harmonics to <-40dBc, thus simplifying the baseband (BB) and LO generation design compared to [5]. The suppression is achieved by controlling the duty cycle of the LO and 1st order RC filtering. Section II discusses multipath system considerations. Section III presents the new architecture; section IV discusses LO duty cycle generation and section V system level simulations.

## II. MULTI PATH TRANSMIITER

Figure 1. shows the basic principle used in [5]. The phase shifts before the nonlinear circuit are assumed to be done in the digital domain for the BB signal, whereas multiphase LO clock drive the mixer switches to provide the second set of phase shifts. After addition the desired signals align in phase, while most of the distortion and harmonics of the PU are cancelled. According to [5] spectral components at $k_{LO}\omega_{LO}+m\omega_{BB}$ are generated by a single-path PU, where $k_{LO}$ is the kth harmonic of the LO frequency $\omega_{LO}$, $m$ is a positive or negative integer, and $\omega_{BB}$ is the single tone BB frequency. For an N-path PU many spectral components can be cancelled, except if [5]:

$$k_{LO} = j \times N + m \qquad \text{where } j = \ldots\text{-2,-1,0,1,2}\ldots \qquad (1)$$

If differential baseband and differential LO signals are used and assuming that the component at $3\omega_{LO}+3\omega_{BB}$ is not dominant, then the first dominant un-cancelled harmonic in an N path PU occurs at N-1 times the LO frequency. The magnitude of this harmonic decreases as N increases, as shown in Figure 2. Even with N=30 the harmonic suppression does not reach -30dBc, while many parallel paths are needed. In this paper we explore options to achieve <-40dBc harmonic suppression by a limited number of paths in combination with simple RC filtering. This is relevant to reduce complexity, as an N-path Transmitter also requires N Digital to Analog Converters (DACs) and N reconstruction filters. Also the

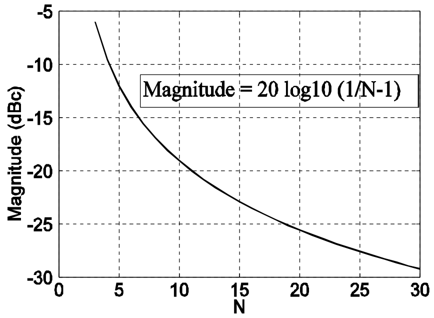



Figure 2. Magnitude of N-1 Harmonic in an N path Transmitter driven by poly-phase 50% duty cycle ideal square-wave LO-signals.

mixers require multiphase clocks and multiphase analog baseband signals to drive them. Multiphase LO-generation becomes more challenging as the number of paths increase, and will require a higher frequency if dividers are used for flexible multiphase LO-generation.

Apart from the LO-phase, duty cycle is another degree of design freedom for LO-generation. In [5] the duty cycle (D) of the LO was made 1/3 to suppress the 3rd LO harmonic. This was done to reduce the $3\omega_{LO}+3\omega_{BB}$ term which is not cancelled by the multipath architecture, as equation (1) is satisfied for m=3 and j=0. The magnitude of the $3\omega_{LO}+3\omega_{BB}$ term relative to the fundamental can be calculated by the relation $(A_p^2 \times a_3 / 3 \times 4a_1)$ [6], assuming D=1/2, where $a_3$ and $a_1$ are the coefficients of the nonlinearity as described by a Taylor approximation [7] and $A_p$ is the amplitude. Assuming $A_p$ to be around 200mV, $a_3/a_1 \approx 3V^{-2}$, this term can be around -40dBc for a switched trans-conductor PU as described in [5]. If some source degeneration is applied for linearization or some RC-filtering is added at the output, this would make this term even lower. So making D=1/3 is not essential in achieving -40dBc suppression of the $3\omega_{LO}+3\omega_{BB}$ term. Due to mismatches and phase inaccuracies, achieving much better harmonic rejection than -40dBc is difficult [5]. Moreover a filter is still needed to reject the first non-cancelled harmonic. In the next section we explore the possibility to achieve -40dBc rejection with a simpler architecture, with less paths allowing for 1st order tunable RC filtering. In contrast to previous work we aim to *suppress all harmonic responses* below -40dBc.

## III. AN 8- PATH TRANSMIITER

The output network of a multipath PU can be implemented as shown in Figure 3a, where $V_o$ represents the mixer output and $R_L$ the antenna load Impedance. If the capacitors (C) are tunable and the inductor (L) value is kept fixed, it is possible to realize a tunable band-pass filter (BPF) response. Below resonance, the frequency response mainly depends on the inductor value (big L for broadband or a small well-defined L for narrowband). Above resonance, RC filtering dominates resulting in a roll-off of 6dB/octave as shown in Figure 3b. The OOB suppression relative to the desired up-converted signal can be increased beyond 6dB/octave if the quality factor Q of the BPF is increased. This in part also depends on how the inductors are implemented. In this work we restrict ourselves to the simpler case with two big inductors L, often

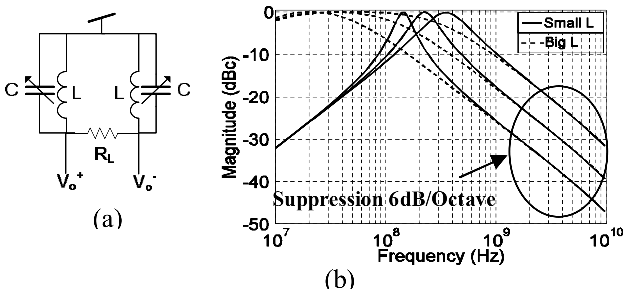



Figure 3 (a) Output Network of multipath PU (b) Tunable BPF response

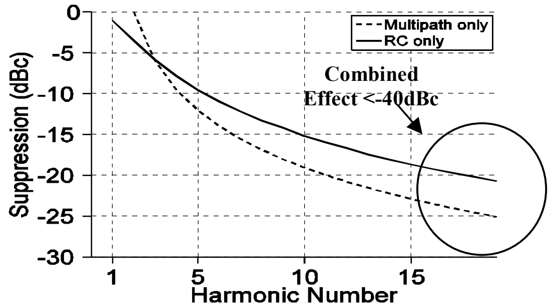


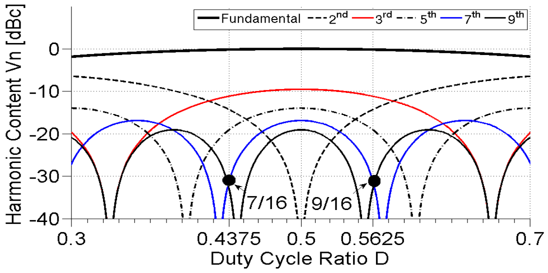

Figure 4 Suppression with 1st order RC Filter and for Multi-path

referred to as "RF chokes". The dominant filtering is now simple low pass filtering due to the fixed $R_L$ and variable C's. The suppression achieved with a simple RC filter is shown in Figure 4 as a function of the LO harmonic numbers. Here the -3dB corner frequency is placed close to the 2nd harmonic, to keep the signal loss for the desired 1st harmonic ≤ 1dB. For convenience the roll-off of Figure 2 is also shown. We see that to achieve <-40dBc suppression in a multipath architecture with simple RC filtering, at-least the 15th harmonic should be the first un-cancelled harmonic, implying N≥ 16. This is the case when the duty cycle D=1/2. Now, consider the possibility to achieve harmonic suppression by varying D of the LO in a multipath architecture. If we keep N even, a 180 degrees shifted version of the LO phase is always available (e.g. for N = 4, phase = 0°, 90°, 180°, 270°). This allows for the suppression of all the even order harmonics, which is desired as the 2nd harmonic becomes dominant when D ≠ ½.We like to keep N as low as possible for complexity reasons, but still aim at <-40dBc suppression. In a 4-path architecture, also referred to as an IQ or image reject mixer, the first few un-cancelled harmonics occur around the 3rd, 5th ,7th harmonic of the LO whereas in a 6-path design they occur around the 5th ,7th,11th harmonic of the LO (see equation (1)). The magnitude of these harmonics is much stronger than -40dBc as seen in Figure 2, and requires significant filtering. An 8-path design presents the opportunity to suppress the two dominant harmonics (7th and 9th) simultaneously, while allowing the next two harmonics (15th and 17th) to be distant enough for simple 1st order RC filtering, as shown in Figure 4. The strength of the 7th and 9th LO harmonics for D=1/2 is -16.9dBc and -19dBc respectively. This can be calculated by taking the magnitude of the Fourier series coefficients of the LO waveform shown in Figure 5b. These coefficients and the amplitude of the nth harmonic $V_n$ are:

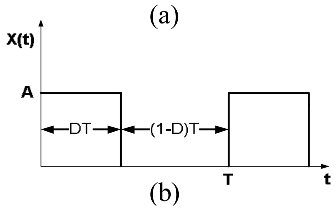
$$a_n = \int_0^T x(t)\cos(n\omega t)dt = \frac{A}{n\cdot\pi}\sin(2\cdot\pi\cdot n\cdot D) \quad (2)$$

$$b_n = \int_0^T x(t)\sin(n\omega t)dt = \frac{A}{n\cdot\pi}(\cos(2\cdot\pi\cdot n\cdot D)-1) \quad (3)$$

$$V_n = \sqrt{a_n^{\ 2}+b_n^{\ 2}} = \frac{A}{n\cdot\pi}\sqrt{2-2\cos(2\cdot\pi\cdot n\cdot D)} \quad (4)$$




(a)




(b)

Figure 5. (a) Harmonic Content as a function of D (b) square wave with duty cycle D.

Table 1. Harmonic Suppression in an 8-path Transmitter

| | Spectral components | $3\omega_{LO}$ $+3\omega_{BB}$ (dBc) | $7\omega_{LO}$ $-\omega_{BB}$ (dBc) | $9\omega_{LO}$ $+\omega_{BB}$ (dBc) | $15\omega_{LO}$ $-\omega_{BB}$ (dBc) | $17\omega_{LO}$ $+\omega_{BB}$ (dBc) |
|---|---|---|---|---|---|---|
| 1 | 50% Duty Cycle | <-38 | -17 | -19 | -24 | -25 |
| 2 | 7/16 or 9/16 Duty Cycle | <-40 | -31 | -33 | -24 | -25 |
| 3 | RC with $BW=2\times\omega_{LO}$ | -4 | -11 | -13 | -17 | -19 |
| 4 | Total Suppression [2+3] | <-44 | -42 | -46 | -41 | -44 |

Equation 4 is plotted as a function of D in Figure 5a, where the desired signal is normalized to 0dB for D=0.5. In the figure it can be seen that both the 7$^{th}$ and 9$^{th}$ LO harmonic are at a level of around -31dBc when D is either 7/16 or 9/16. These are the only two points in Figure 5a where *both* the 7$^{th}$ and 9$^{th}$ harmonic components are below -31dBc *simultaneously*.

This represents an improvement in harmonic suppression of at least 14 dB as compared to the case when D=1/2. As the 7$^{th}$ harmonic is more than 2 octaves from the fundamental, a simple first order filter can implement the missing 10dB to achieve <-40dBc (2 octaves ⇔ 2x6dB=12dB filtering). The next two most dominant harmonics for an 8-path transmitter occur at 15 and 17 times the LO, which are at a level of around -24dBc. A first order filter can have around -20dB suppression there, which again brings the suppression at < -40dBc. Higher harmonics are at an even lower amplitude level. A summary of the amplitude of the dominant spectral components in an 8-path transmitter is given in Table 1. Here -3dB corner of the RC-filter is placed around the 2$^{nd}$ harmonic of the LO. A design based on harmonic rejection mixing to suppress higher LO-harmonics has also been proposed in [3], but with active RF filters which can be undesired for nonlinearity and power efficiency reasons.

One might wonder what happens for less paths. A 4 path design is the minimum even N design with image rejection, required for single sideband transmission. In order to allow the 15$^{th}$ harmonic to be the first un-cancelled harmonic which can be <-40dBc (Figure 4), the 3$^{rd}$, 5$^{th}$…13$^{th}$ harmonic would need to be significantly suppressed simultaneously, which is not possible with any single D value. A similar argument can be given for not choosing 6 paths. Therefore the next even N of 8 is chosen.

## IV. LO GENERATION

An 8-path transmitter requires 8-phase LO signals to drive the mixer switches. The LO phases can be generated by a divide-by-8 circuit [8]. For low phase mismatch it is desired to drive all flip-flops (FF) in a divide-by-8 circuit by the same master clock. A chain of 8 FF cells is needed to generate all the phases. However to make the duty cycle ratio 7/16 requires the implementation of 8 additional FF cells (for divide by 16). This is possible but requires extra power consumption and also limits the maximum output frequency in a particular process technology, since the input frequency of the divider chain then has to be 16 times higher than the LO frequency.. Thus an 8-phase LO generation circuit is preferred over a 16-phase circuit.

In a divide-by-8 circuit a duty cycle ratio of k/8 can easily be implemented, where k =1, 2...7. For power efficiency reasons it is best to operate the mixers in the region below D=0.5, since for an ideal 8-path Transmitter, the LO harmonic content for D=0.4375 (7/16) is the same for D=0.5625 (9/16) while the power consumed is 7/9 times less, as the mixers remain ON for less time. In an 8-phase LO generation circuit a duty cycle ratio of 3/8 and 4/8 represent the two nearest points to the desired ratio of 7/16. A simple way to reach this ratio is to make the LO edges slower. One possibility is shown in Figure 6. Starting from a duty cycle ratio of 3/8 the falling edge can be slowed. This way the effective LO waveform has un-symmetric rise and fall times but the effective time the mixer switch remains ON can be increased, thus approximating a 7/16 ratio. If the LO waveform switches between 0 and VDD and the mixer switch turns on at VDD/2, the required edge delay as a ratio of the LO period is 1/8. For a 1GHz LO, this means a fall time of 125ps.

For a frequency agile transmitter making the falling edge delayed by a ratio of 1/8 of the LO period over a wide frequency range requires adding a tuning mechanism to the LO waveform. One possibility can be to digitally update the size of the LO inverter buffers to provide the necessary fall time. For example, in a CMOS inverter the ratio of the size of the NMOS and PMOS, can be adjusted such that it gives un-symmetric rise and fall times. Enough digital control is required in each path, so that the delay ratio of 1/8 can be

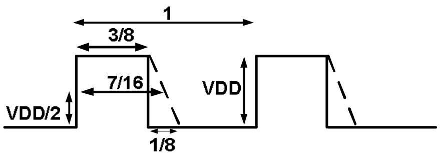



Figure 6. Making D = 7/16 from D = 3/8

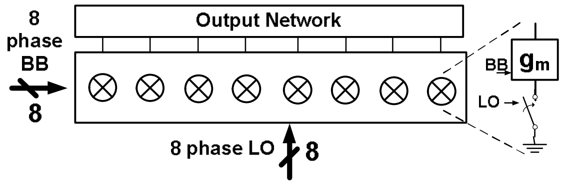



Figure 7. 8-Path Power Up-converter Architecture

achieved over a wide bandwidth and hence the required suppression. This form of digital control can also help in tuning out possible LO phase-mismatch occurring due to process spread.

## V. SIMULATIONS

System level simulations were performed in MATLAB to verify the harmonic suppression achievable in an 8-path up-converter shown in Figure 7. As the system concept relies on timing, we also examine its sensitivity to timing errors. The $g_m$ block provides the gain and V-I conversion, while the switch provides the up-conversion. The output network contains tunable RC filtering with a fixed L. The multiphase single tone baseband and LO signals were first idealized. The LO duty cycle was set at D= 7/16. The output spectrum of the transmitter without filter is shown in Figure 8a. Then, after applying RC filtering all the harmonics are suppressed to <-40dBc, as seen in Figure 8b. A bank of switched capacitors is needed to cover a wide frequency range and also mismatches due to PVT variations. As a cognitive radio would have a spectrum analyzer on-board, we think it is realistic to sense spectral components and provide feedback to the transmitter to tune the duty cycle and filter appropriately. Further work is needed to show this in practice.

In order to assess the sensitivity to phase and amplitude inaccuracies, a 1% random phase mismatch was introduced in the LO paths and also the magnitude of the coefficients of the nonlinearity ($a_1$,$a_3$) were randomly varied by 1%. Simulation results for 50 Monte Carlo runs are given in Figure 9 after RC filtering. The variation is less than 0.5dB from the nominal suppression for the un-cancelled harmonics (occurring around $7^{th}$, $9^{th}$, $15^{th}$, $17^{th}$ harmonics). This relatively small variation results because the harmonic contributions from all the paths add up almost in phase [5]. Much more dB variation is observed in harmonics which are nominally completely cancelled, but also their strength stays below -42dBc, except for the LO leakage (~ -37dBc). For a direct conversion transmitter LO leakage is in-band, so it does not cause OOB emissions and only degrades the in-band distortion slightly.

## VI. CONCLUSION

A compact poly-phase multipath transmitter system concept is presented for cognitive radio. The main advantage is that *all* harmonic responses are suppressed below -40dBc even for just 8-paths. The trade-off made compared to higher number of paths is less suppression for all harmonics instead of more suppression for a few harmonics. The suppression is achieved through a combination of 8-paths, duty cycle control and $1^{st}$ order passive filtering. This simplifies the baseband design where 8 DACs (or 4 differential DACs) can provide the necessary multiphase baseband signals. The LO generation can also be simplified and the necessary duty cycle can be generated from an 8 phase LO-generation circuit. Compared to the baseband and LO generation requirements of an 18 path design [5], this design can be more area and power efficient, with only slightly less harmonic suppression.

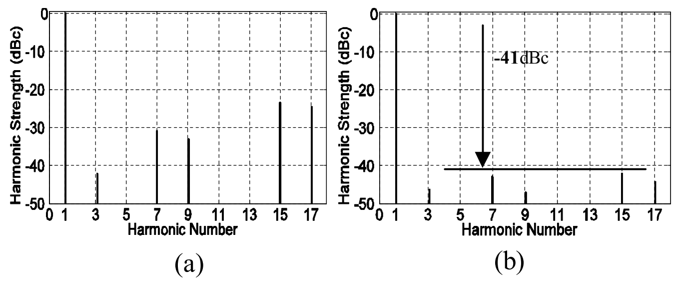



Figure 8. Output Spectrum (a) before RC filtering (b) after RC filtering

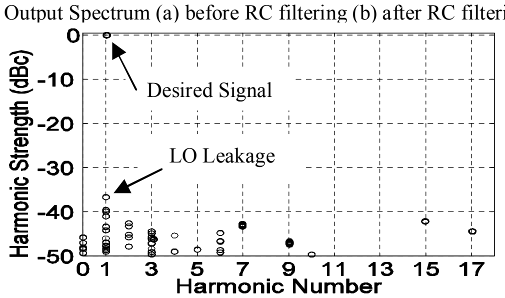



Figure 9 Effect of phase and amplitude mismatch on an 8-path Transmitter

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