Digitally Assisted Wideband Compensation of Parallel RF Signal Paths in a Transmitter

Prasidh Ramabadran[1], Sidath Madhuwantha[1], Pavel Afanasyev[1], Ronan Farrell[1], Lazaro Marco[2], Sergio Pires[2] and John Dooley[1]
[1]National University of Ireland, Maynooth, Ireland, [2]Ampleon, Netherlands

Abstract — **Modern wireless communication equipment such as outphasing power amplifiers or systems like massive-MIMO rely heavily on transmission of complex wideband modulated radio frequency signals on parallel signal paths. As these signal bandwidths increase, wireless transmitters are more susceptible to amplitude and phase distortions across frequency. We propose a novel method to quantify the complex signal distortions in each transmit path and a technique to pre-compensate the transmitter over a wide bandwidth of interest. This work has been experimentally validated with measured results on two separate RF test benches using signal bandwidths up to 100 MHz. An outphasing power amplifier bench for WCDMA at S band requiring 4 signal paths and a satellite uplink modulator using 8-PSK at Ku band requiring two signal paths were tested in the experimental validation. Further, it is also validated that this method requires only one iteration to calibrate a set of parallel RF signal paths.**

***Index Terms*-RF/digital mixed-signal measurement and calibration**

I. INTRODUCTION

Modern wireless hardware employs digital modulation schemes where the digital data to be transmitted is modulated on RF carrier signals. In communication systems such as beamforming or massive MIMO there are multiple signal paths where the processing for each path takes place in parallel stages. The relative gain and phase relations between all paths over the desired bandwidth of operation are critical to operation. Another example of a system with parallel complex/vector modulated transmit paths is an outphasing power amplifier in which the real and imaginary parts of a digital baseband are modulated on to RF carrier sinusoids of equal amplitudes and quadrature in phase by means of signal multiplication using an RF mixer in each path. The outputs of the two parallel modulators in quadrature are combined to yield a vector modulated RF signal whose values for amplitude and phase act as identifiers for the digital information/symbol being transmitted. The timing, gain, phase and the imbalances between the two parallel paths and their distortions in each path over the required bandwidth will alter the amplitudes and phases of the modulated RF carrier which critically affect the performance of the final power amplification stage.

The task of achieving amplitude and phase balance between two paths over wide bandwidths with single, multi-tone stimuli have previously been accomplished [1]-[3]. However, this is done using iterative procedures which can prove time consuming. [4,5] illustrate the principles of calibrating dc and timing offsets and wide bandwidth amplitude, phase distortions between RF signal paths using even spaced sinusoidal tones. [6] applies a chirp stimulus to calibrate the wideband response of the paths in a phased array system through time reversal techniques. We propose to calibrate the timing, gain and phase offsets, wideband gain roll off and group delay distortions in a non-iterative manner.

The remainder of this paper is arranged as follows: In section II we describe the proposed wideband stimulus generation and extraction of calibration coefficients to pre-compensate multiple parallel RF signal paths. The experimentally measured results to validate the technique are presented in section III. Finally in section IV we cover the main conclusions from this work.

II. THE PROPOSED TECHNIQUE

The proposed technique involves a combination of time and frequency domain analyses of the impairments and imbalances that a wideband test signal through a super-heterodyne or a homodyne transmitter. A matrix of coefficients is subsequently generated in the form of an FIR filter kernel similar to [4] to enable the pre-compensation of signal impairments.

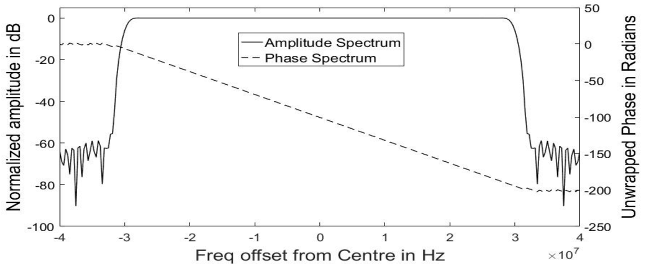

Figure 1 Spectrum of the test signal (stimulus) for quasi linear or weakly non-linear modules.

2.1 Calibration of Linear and Weakly Non-Linear transmit paths.

2.1.1 Choice of the stimulus: A desirable test signal (stimulus) is one that excites a band of desired frequencies with a reference amplitude, maintains frequency components outside the band of interest at reasonably low levels and maintains a pre-determined phase profile across the band of interest. A practical choice for this requirement is a truncated sinc pulse whose spectral characteristics exhibit a near flat amplitude and a linear phase profile across a wide bandwidth as shown in

978-1-5386-5450-7/18/$31.00

Figure 1. A popular method to implement digital filters is to use windowed sinc kernels such as ‘Hamming Window’ defined by the equation

$$h(n) = sin\frac{\left(2\pi f_c\left(n-\frac{M}{2}\right)\right)}{\left(n-\frac{M}{2}\right)}\left(0.54-0.46\left(\frac{cos(2\pi n)}{M}\right)\right) \quad (1)$$

Where fc is half of the desired bandwidth, M is the sample length of the sinc pulse, n is the sample number ranging from 0 to M whose value is selected depending upon the desired transition region between the pass and stop bands. Here the equation is used to generate a stimulus signal instead of a filter kernel. M can be varied to tune the peak to average power ratio (PAPR) of the stimulus depending upon the signal power limits of the hardware under test.

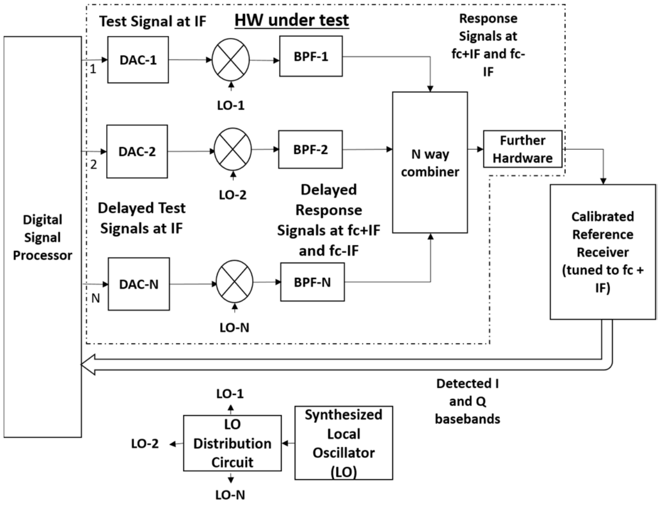

Figure 2. Test set up for an N Path modulated RF transmitter

2.1.2 Calibrating a single RF path: The test signal described above is generated digitally by the DSP either at an intermediate frequency (IF) or Zero IF, converted to analogue form by the Digital to Analogue Converter (DAC) and applied to the RF signal path to be compensated. Its response in the band of interest is captured by a calibrated receiver such as a vector signal analyzer (VSA) tuned to the intended center frequency. The fast Fourier transform (FFT) of the response of the signal path under test is computed to obtain the amplitude and phase response at each excited frequency. The number of FFT points can be chosen depending upon the resolution needed. For the convenience of subtraction, the amplitude spectrum of the FFT is plotted on dB scale. The amplitude spectrum computed directly indicates the in-band amplitude ripple and offset across frequency. The amplitudes are normalized to the maximum amplitude and directly subtracted from the normalized amplitude spectrum of the stimulus to yield the amplitude correction factors over the band of interest. The first derivative of the phase spectrum in the band of interest is computed numerically to yield the in-band group delay variation or phase distortion over the bandwidth.

$$Tg = \left(\frac{\boldsymbol{d\theta}}{\boldsymbol{d\omega}}\right) \quad (2)$$

Where Tg is the group delay is in seconds, θ is the phase in radians and ω is the angular frequency in radians/second. The first numerical derivative of the phase spectrum of the response is subtracted from that of the stimulus to yield the phase correction factors over the band of interest.

The amplitude correction factors converted to linear scale and phase correction factors obtained for the path represent the frequency domain amplitudes and phases of a windowing pre-compensation filter for the respective path whose inverse FFT (IFFT) would form a filter kernel that could be convolved with the respective time domain data signals. A calibrated heterodyne receiver consisting of a scalar down-converter and an Analogue to Digital converter (ADC) with the required bandwidth to capture the response may also be used instead of a VSA. The digitized down-converted signal may be vector demodulated digitally to obtain the amplitude and phase of the responses over the desired bandwidth. If the transmitter under test resides in a transceiver, its own calibrated receiver section may also be used for vector demodulation.

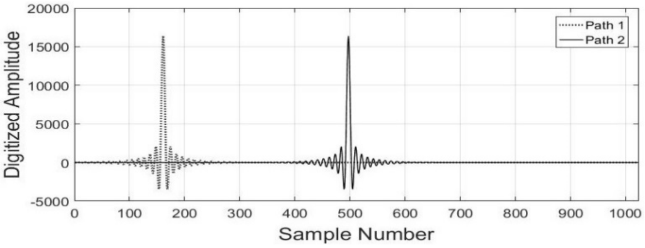

Figure 3 Digitized Stimuli generated for the two paths before up-conversion

2.1.3 Calibrating multiple RF paths: This involves application of the above test signal to the parallel RF signal processing paths at unique non-overlapping time intervals and evaluating the response to yield the correction factors for each path and relative variations between them. For example, consider a case of calibrating two parallel RF paths in the block diagram of the N path modulated transmitter shown in Figure 2, where paths ‘1’ and ‘2’ are expected to be in phase quadrature along with equal amplitude and uniform phase responses across the intended band of operation. A digital signal processor generates the digital baseband signal at the intended IF and is applied to the DAC in each path whose outputs are further mixed with quadrature phases of the signal generated by local oscillator LO at frequency fc and combined to yield a vector modulated RF output. Consider a digitally generated test signal i.e. a truncated sinc pulse of 320 samples. This signal is zero-padded to extend the number of samples to the next power of 2 i.e. 1024 samples to yield the stimulus to the first path. A power of 2 is selected as the total length of the signal for the ease of computation of FFT. The stimulus to the second path is obtained by circularly delaying the stimulus of the first path such that it remains zero-valued in the duration when the stimulus to the first path takes the values of the truncated sinc pulse plus some guard interval. Two such signals are plotted on the same axis in Figure 3. The response captured by the reference receiver at the center frequency *fc* is the sum of the responses of the individual paths which are complex values representing the amplitude and phase at each sampling interval. Considering the response signal captured in this example, samples from 1 to 336 represent the response of RF path 1 and samples from 337 to 583 represent the second RF path, path 2. The amplitudes of the extracted individual responses can then be cross-correlated to determine the relative timing offset between the two paths in terms of number of sampling intervals and the relative delay. It is for this reason that it is possible using this proposed method to derive the calibration co-efficients for all parallel signal paths in a single iteration.

Path 1 or 2 could be considered the reference path. The amplitude and phase correction factors for the reference path are obtained as detailed in section 2.2. For the remaining path, it is intended to correct its own impairments and also establish the desired amplitude balance and quadrature phase relation relative to the reference path. To achieve this, the amplitudes of the FFT of the non-reference path are normalized to the maximum amplitude value of the FFT of the reference path and then subject to the calibration process detailed in section 2.1.2. The mean of the differences of phase values between the responses of the reference and non-reference paths in the band of interest is computed. This is the bandwidth independent phase difference caused primarily by the difference in the LO phases fed to the two paths. This should ideally be equal to $\frac{\pi}{2}$ radians if the two paths are in quadrature. If different, its deviation relative to $\frac{\pi}{2}$ radians is added as an offset to the frequency domain phase correction factors computed for the non-reference path. With the frequency domain amplitude and phase correction factors in hand for each path, the pre-compensation filter kernel for each path can be obtained as explained in section 2.1.2. The relative gain offset is determined by taking the ratio of the Root Mean Square (RMS) value of the time domain response obtained for the reference path to that of the non-reference path. The reference path signal is delayed by the required number of samples to compensate the impact of timing offset between the two paths. As seen above, the advantage of the proposed technique is that it needs just one iteration and just one measurement operation with the reference receiver to accomplish the required calibration of a set of parallel signal paths. The technique is scalable to a larger number of parallel transmitter paths with ease which only requires applying further delayed versions of the stimulus to each path and calibrating them as detailed above.

2.2 Calibration of Strongly Non- Linear Modules:

Despite its desirable spectral characteristics, a truncated sinc pulse would lead to gain compression when applied to strongly non-linear modules such as Class C amplifiers which often require the stimulus signals to have constant envelope. An alternate choice proposed is a linearly swept Frequency Modulated (FM) Chirp pulse which could be viewed as a sine wave whose frequency varies linearly from a minimum to a maximum value at every sample. The spectrum of this signal does not have a flat amplitude profile like the sinc pulse and exhibits peaks depending on the rate of sweep of frequency. The significant bandwidth of an FM signal may be estimated as twice the sum of the modulating frequency and frequency deviation. Generating a Chirp signal whose frequency varies linearly from 0 to 45 MHz over a period of 25μsec corresponds to an FM signal with a frequency deviation of 45 MHz and modulating frequency of 40 kHz which would have a significant bandwidth of 90.08MHz. A digitally generated chirp signal is described by the following equations.

$$y(n) = \sin\left\{2\pi\left(\frac{knt_s}{2} + \mathrm{f_{min}}\right)\mathrm{nt_s}\right\} \tag{3}$$

$$k = (f_{max} - f_{min})/M \tag{4}$$

Where $y(n)$ is the value of the nth sample, t_s is the sampling interval, M is the total number of samples, k is the frequency variation parameter, f_{max} and f_{min} are the maximum and the minimum frequencies respectively. The spectrum of this signal is as shown in Figure 4. The time domain plot would resemble a frequency modulated signal with constant envelope.

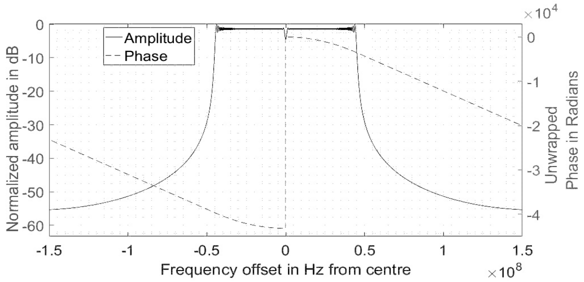

Figure 4 Spectrum of the stimulus for characterizing strongly nonlinear modules.

The calibration procedure is the same as that described in sections 2.1.2 and 2.1.3 except that the correction co-efficient matrix extracted in the frequency domain needs to be filtered by a windowing function such as a Hamming Window to limit the response to the band of interest, owing to the relatively wide transition region between the pass and stop bands in the stimulus and response signals. The chirp pulse technique is marginally more computationally demanding relative to the sinc pulse characterization and compensation technique.

III. EXPERIMENTAL VALIDATION

The proposed technique was validated using a DAC34SH84, a 4-channel 16-bit DAC, from Texas Instruments followed by a pair of TRF3705 quadrature up-converters (QUC). Channels '1-2' and '3-4' of the DAC together with the QUCs formed the first and second modulator quadrature pairs. The VSA FSQ 40 from Rohde and Schwarz with a vector demodulation bandwidth of 120 MHz was used as the reference receiver.

Case 1 - Zero IF with 4 paths: The 4 paths of LINC transmitter's modulator built with the quadrature modulator pairs described above were calibrated at zero IF over a bandwidth of 100 MHz. LINC baseband signals were generated for a 5 MHz WCDMA baseband signal with a power crest factor of 7dB. The LINC signals occupied a bandwidth over 100 MHz due to bandwidth expansion in the resulting phase modulated signals with high modulation index. Phase errors were deliberately introduced using LO feed cables of non-uniform lengths between the paths to replicate a more challenging use case. The calibration was performed using both chirp and truncated sinc stimuli in two different trials at a sample rate of 307.2 MSPS, center frequency of 2.24 GHz at a bandwidth of 100 MHz with 1024 FFT points. An improvement in Adjacent Channel Power Rejection (ACPR) of 36 dB was observed as shown in Figure 5.

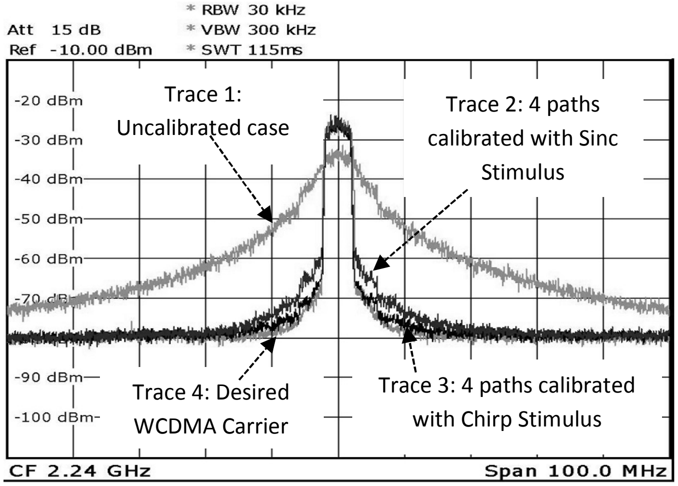

Figure 5 Spectra of the LINC signal generated with calibrated and uncalibrated signal paths at 2.24 GHz captured on R&S FSL.

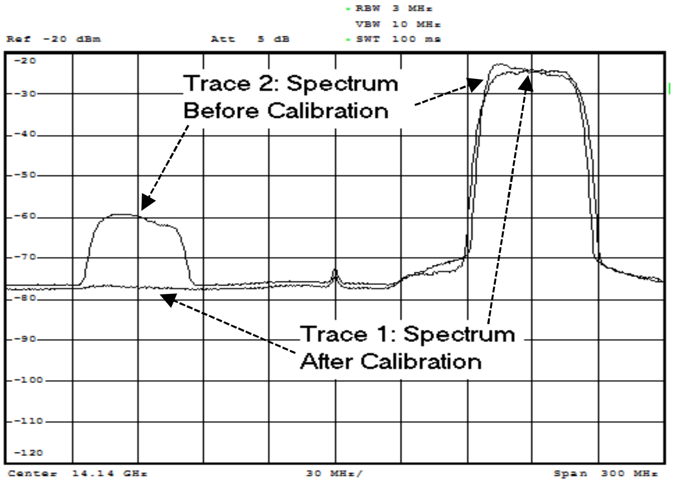

Figure 6 Spectra of 8PSK signal before and after signal path calibration captured on R&S FSQ. Centre = 14.14 GHz, Span = 300 MHz, Ref = -20 dBm, RBW = 3 MHz

Case 2 -Double Heterodyne, 2 paths: For the second validation test an 8PSK modulator at S-Band built with the first quadrature pair described above, up-converted to Ku Band was used with the following settings: Symbol rate of 40MSPS, Root Raised Cosine Filter roll off factor of 0.2, digital IF generated at 90 MHz. The LO at S-Band was set at 2.24 GHz yielding up-converted signals at 2.33 GHz (desired) and 2.15 GHz (image). These were up-converted to frequencies of 14.23 GHz (wanted) and 14.05 GHz (image) respectively by a Ku band up-converter fed with an LO at frequency 11.9 GHz. A cavity tuned filter with a pass band of 14 to 14.5 GHz was used at the output of the Ku Band mixer. This second up-converter and the cavity tuned filter are represented by the block 'Further Hardware' in Figure 2. As the system here is weakly non-linear, a calibration using sinc stimulus would suffice. In this case two path calibration was performed on the entire cascaded system at 14.23 GHz by applying a truncated sinc stimulus of bandwidth 60 MHz and 512 points FFT for analysis. An improvement in image rejection by 17 dB was observed as in Figure 6 along with an improvement in Modulation Error Rate (MER) by approximately 7 dB i.e. from 16.9 dB to 23.8 dB.

IV. CONCLUSION

The proposed technique provides a low complexity non-iterative method to calibrate the timing, amplitude and phase offsets and distortions across multiple RF signal paths over a wide frequency band of interest and the operation requires no more than one iteration per set of parallel paths. The technique is scalable with number of paths and applicable even if the desired amplitude and phase relations between RF paths under test are set arbitrarily.

ACKNOWLEDGEMENT

This paper has emanated from research conducted with the financial support of Science Foundation Ireland (SFI) and is co-funded under the European Regional Development Fund under Grant Number 13/RC/2077." We also acknowledge the support provided by Mr. James Kinsella of NUI, Maynooth and Dr Bill O'Brien of Non Linear Systems, Ireland in this project.

REFERENCES

[1] A Rezola, JF Sevillano, I Gurutzeaga, D Rio, R Berenguer, I Velez, "Built-in-Self-Calibration for IQ imbalance in Wideband Millimeter-Wave Gigabit Transmitters," *IEEE Trans. Microw. Theory Tech.*, vol. 65, no. 11, Nov. 2017.

[2] A Khandelwal, A Verma, "A Novel Gain, Phase and Offset Calibration Scheme for Wideband Direct-Conversion Transmitters", Vehicular Technology Conference (VTC Spring), 2015 IEEE 81st, DOI: 10.1109/VTCSpring.2015.7145633.

[3] Z Zhu, X Huang, "An Iterative Calibration Technique for LINC Transmitter" Circuits and Systems (MWSCAS), 2013 IEEE 56th International Midwest Symposium on, DOI: 10.1109/MWSCAS.2013.6674666

[4] Keysight | Calibration Techniques for Improved MIMO and Beamsteering Characterization - White Paper

[5] Keysight | Enhancing Measurement Performance for the Testing of Wideband MIMO Signals- White Paper

[6] M. Longbrake *et al.*, "Wideband phased array calibration method for digital beamforming," *2012 IEEE National Aerospace and Electronics Conference (NAECON)*, Dayton, OH, 2012, pp. 11-17. DOI: 10.1109/NAECON.2012.6531016