DIGITAL PREDISTORTION OF CONCURRENT MULTIBAND COMMUNICATION SYSTEMS

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ABSTRACT

In this semi-tutorial paper, we derive the general form of a predistorter for a concurrent multiband system. Further, we derive complexity-reduced forms, and we compare the resulting algorithms with previously proposed algorithms in the literature.

1. INTRODUCTION

In wireless communication transmitters, high power-efficiency is critical, both in central nodes (base stations) and in battery-powered devices. This makes it necessary to drive the power amplifier (PA) close to its saturation region, which results in unwanted spectral emissions and in-band distortion. A common remedy used today today is to apply digital predistortion (DPD), where a pre-inverse to the amplifier is computed with a digital baseband algorithm.

Lately, the need to accommodate simultaneous transmission of multi-band signals through a single PA has increased, and research on PAs supporting multiple bands has been intense [1, 2] . As in single-band PAs, the nonlinear distortion is an important problem, and DPD will be necessary. However, predistortion of multiple signals at separate frequencies, simultaneously amplified in a single PA, is quite different from the single-band problem. Severe cross-dependency between the two signals leads to a more difficult problem. The problem has been addressed by Bassam et al. in [3-5], where a dual-input DPD (2D-DPD) structure was proposed based on a memory-polynomial [6] description of the amplifier. Liu et al. [7, 8] continued this work with complexity reduction of the original 2D-DPD model, by allowing for some lost accuracy. Other reports attacking the same problem are e.g. [9] where the Volterra-DDR [10] is reworked for concurrent dual-band operation, [11] where look-up tables are used, and [12] where time misalignment between the two bands are specifically addressed. There are also work that addresses the general problem of dual-input linearization [13], which describes a general dual-input DPD scheme, not dedicated to concurrent dual-band operation.

While the above reports have made good progress in solving the concurrent dual-band problem with low-complex DPD algorithms, there have been no reports on the general form of such a predistorter. Most previous reports are based on a memory-polynomial description of the amplifier, which may not be accurate enough, particularly if the amplifier is wideband and the DPD must utilize cross-terms in the description.

The main contribution of this paper is that we develop the general form of a concurrent dual-band DPD, based on a general Volterra description of the amplifier. We illustrate which cross-terms that must be included for a full description, and which cross-terms that can be disregarded. Then, we derive complexity-reduced versions of the general equation, and we show that previous techniques can be written by excluding some cross-terms from the original form. Finally, we show the performance of the developed techniques, and compare the performance with previously proposed algorithms.

2. SYSTEM MODEL

2.1 The concurrent multiband input problem

A block diagram of the intended system is given in Fig. 1.

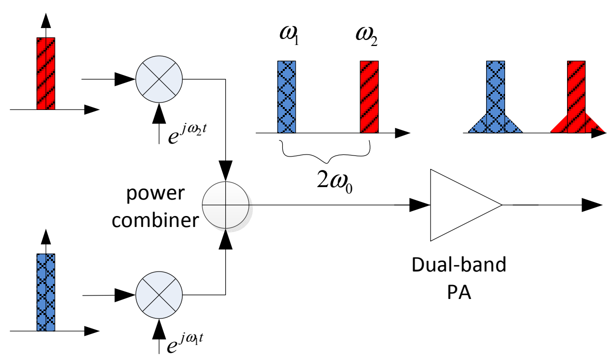

Fig.1. Block diagram of a concurrent dual-band system.

Two narrowband signals, at widely spaced frequencies, are fed to a single wideband amplifier. It is assumed that the harmonics of the two signals are not overlapping.

Now we wish to design a digital predistorter (DPD) for the system. Since the two signals will interfere with each other when fed through the nonlinear PA, we will need to use predistorters with knowledge of both input signals. In the following section, we will derive a dual-input predistorter for this purpose.

2.2 The general discrete complex-baseband model

The general form of the baseband Volterra DPD is given in (1).

$$
\begin{aligned}
y_n &= \sum_{m_1=0}^{M} \lambda_{m_1}^{(1)} x_{n-m_1} + \sum_{m_1=0}^{M} \sum_{m_2=m_1}^{M} \sum_{m_3=0}^{M} \lambda_{m_1 m_2 m_3}^{(3)} \underbrace{x_{n-m_1} x_{n-m_2} x_{n-m_3}^{*}}_{\text{3rd order monomial}} \\
&+ \sum_{m_1=0}^{M} \sum_{m_2=m_1}^{M} \sum_{m_3=m_2}^{M} \sum_{m_4=0}^{M} \sum_{m_5=m_4}^{M} \lambda_{m_1 m_2 m_3 m_4 m_5}^{(5)} \underbrace{x_{n-m_1} x_{n-m_2} x_{n-m_3} x_{n-m_4}^{*} x_{n-m_5}^{*}}_{\text{5th order monomial}}
\end{aligned}
\tag{1}
$$

As shown in e.g. [12], this Volterra series is an accurate truncated time-discrete baseband description of any nonlinear RF predistorter. Note that the monomials of the Volterra are all of odd order, and that they consist of the product of delayed inputs and their complex-conjugates, always with one less complex-conjugate factor. Other monomials, with even orders and with other proportions of complex-conjugated factors, contains signals that falls outside the interesting bands and will be filtered away.

The sampling frequency of the signals must be at least twice the bandwidth of the output signal y_n, which will be *p* times the bandwidth of the input signal x_n (*p* is the nonlinear order). Particularly for dual-band operation this can be prohibitively high, and in the following we will show how the sampling frequency can be lowered.

2.3 Definition of the input signal

In the concurrent dual-band setting, the input signal consists of two narrow-band signals $x_{1,\mathrm{n}}$ and $x_{2,\mathrm{n}}$ at different carrier frequencies. Such a signal can, in baseband, be described as

$$x_n = x_{1,\mathrm{n}} e^{j\omega_0 n} + x_{2,\mathrm{n}} e^{-j\omega_0 n}. \tag{2}$$

where we have centered the baseband to the average of the signal frequencies $(\omega_1 + \omega_2)/2$. A requirement of the derivations in the following section is that the frequency distance between the two signals is large enough to avoid overlap between the spectra after the nonlinearity. In practice, this means that $\omega_0 \geq 5W$ or so, with *W* being the signal bandwidth. If the separation between the signals is smaller than that, it is probably better to avoid concurrent dual-band linearization altogether, and linearize the combined signal x_n directly.

3. CONCURRENT DUAL-INPUT VOLTERRA

We will now derive the general form of the Volterra series when the input signal is given by (2). We are interested in the frequency region around ω_0, and will exclude other frequencies from the expression.

3.1

1st order (linear) terms: Inserting (2) in (1) yield the following linear term

$$x_{n-m_1} = \boxed{x_{1,n-m_1} e^{j\omega_0 (n-m_1)}} + x_{2,n-m_1} e^{-j\omega_0 (n-m_1)}$$

The grey box in the equation indicates the interesting frequency band around ω_0; the other terms can be filtered away.

3rd order **terms:** All 3rd order terms are of the form $x_{n-m_1} x_{n-m_2} x_{n-m_3}^{*}$. Inserting (2), we have

$$
\begin{aligned}
& x_{n-m_1} x_{n-m_2} x_{n-m_3}^{*} = \left(x_{1,n-m_1} e^{j\omega_0 (n-m_1)} + x_{2,n-m_1} e^{-j\omega_0 (n-m_1)} \right) \cdot \\
& \left(x_{1,n-m_2} e^{j\omega_0 (n-m_2)} + x_{2,n-m_2} e^{-j\omega_0 (n-m_2)} \right) \left(x_{1,n-m_3}^{*} e^{-j\omega_0 (n-m_3)} + x_{2,n-m_3}^{*} e^{j\omega_0 (n-m_3)} \right) \\
& = e^{-3j\omega_0 n} x_{2,n-m_1} x_{2,n-m_2} x_{1,n-m_3}^{*} e^{j(m_1+m_2+m_3)} \\
& + e^{-j\omega_0 n} \Big(x_{2,n-m_1} x_{2,n-m_2} x_{2,n-m_3}^{*} e^{j(m_1+m_2-m_3)} + x_{2,n-m_1} x_{1,n-m_2} x_{1,n-m_3}^{*} e^{j(m_1-m_2+m_3)} \\
& \qquad + x_{1,n-m_1} x_{2,n-m_2} x_{1,n-m_3}^{*} e^{j(-m_1+m_2+m_3)} \Big) \\
& \boxed{\begin{aligned} + e^{j\omega_0 n} \Big(& x_{1,n-m_1} x_{1,n-m_2} x_{1,n-m_3}^{*} e^{j(-m_1-m_2+m_3)} + x_{1,n-m_1} x_{2,n-m_2} x_{2,n-m_3}^{*} e^{j(-m_1+m_2-m_3)} \\ & + x_{2,n-m_1} x_{1,n-m_2} x_{2,n-m_3}^{*} e^{j(m_1-m_2-m_3)} \Big) \end{aligned}} \\
& + e^{3j\omega_0 n} x_{1,n-m_1} x_{1,n-m_2} x_{2,n-m_3}^{*} e^{j(-m_1-m_2-m_3)}
\end{aligned}
$$

Again, we are only interested in the frequency region around ω_0, which is indicated by the grey box above.

$$
\begin{aligned}
y_n &= \sum_{m_1=0}^{M} a_{m_1}^{(1)} x_{1,n-m_1} \\
&+ \sum_{m_1=0}^{M} \sum_{m_2=m_1}^{M} \sum_{m_3=0}^{M} a_{m_1 m_2 m_3}^{(3)} \left(x_{1,n-m_1} x_{1,n-m_2} x_{1,n-m_3}^{*} e^{j(-m_1-m_2+m_3)} + x_{1,n-m_1} x_{2,n-m_2} x_{2,n-m_3}^{*} e^{j(-m_1+m_2-m_3)} + x_{2,n-m_1} x_{1,n-m_2} x_{2,n-m_3}^{*} e^{j(m_1-m_2-m_3)} \right) \\
&+ \sum_{m_1=0}^{M} \sum_{m_2=m_1}^{M} \sum_{m_3=m_2}^{M} \sum_{m_4=0}^{M} \sum_{m_5=m_4}^{M} a_{m_1 m_2 m_3 m_4 m_5}^{(5)} \Big(x_{1,n-m_1} x_{1,n-m_2} x_{1,n-m_3} x_{1,n-m_4}^{*} x_{1,n-m_5}^{*} e^{j\omega_0 (-m_1-m_2-m_3+m_4+m_5)} + x_{1,n-m_1} x_{1,n-m_2} x_{2,n-m_3} x_{2,n-m_4}^{*} x_{1,n-m_5}^{*} e^{j\omega_0 (-m_1-m_2+m_3-m_4+m_5)} \\
&\qquad + x_{1,n-m_1} x_{1,n-m_2} x_{2,n-m_3} x_{1,n-m_4}^{*} x_{2,n-m_5}^{*} e^{j\omega_0 (-m_1-m_2+m_3+m_4-m_5)} + x_{1,n-m_1} x_{2,n-m_2} x_{1,n-m_3} x_{2,n-m_4}^{*} x_{1,n-m_5}^{*} e^{j\omega_0 (-m_1+m_2-m_3-m_4+m_5)} \\
&\qquad + x_{1,n-m_1} x_{2,n-m_2} x_{1,n-m_3} x_{1,n-m_4}^{*} x_{2,n-m_5}^{*} e^{j\omega_0 (-m_1+m_2-m_3+m_4-m_5)} + x_{1,n-m_1} x_{2,n-m_2} x_{2,n-m_3} x_{2,n-m_4}^{*} x_{2,n-m_5}^{*} e^{j\omega_0 (-m_1+m_2+m_3-m_4-m_5)} \\
&\qquad + x_{2,n-m_1} x_{1,n-m_2} x_{1,n-m_3} x_{2,n-m_4}^{*} x_{1,n-m_5}^{*} e^{j\omega_0 (m_1-m_2-m_3-m_4+m_5)} + x_{2,n-m_1} x_{1,n-m_2} x_{1,n-m_3} x_{1,n-m_4}^{*} x_{2,n-m_5}^{*} e^{j\omega_0 (m_1-m_2-m_3+m_4-m_5)} \\
&\qquad + x_{2,n-m_1} x_{1,n-m_2} x_{2,n-m_3} x_{2,n-m_4}^{*} x_{2,n-m_5}^{*} e^{j\omega_0 (m_1-m_2+m_3-m_4-m_5)} + x_{2,n-m_1} x_{2,n-m_2} x_{1,n-m_3} x_{2,n-m_4}^{*} x_{2,n-m_5}^{*} e^{j\omega_0 (m_1+m_2-m_3-m_4-m_5)} \Big)
\end{aligned}
$$

5th and higher order terms: The higher order terms get increasingly complicated. The general technique can, however, be understood from the complex exponential in the expression for order 3. We see that there are always 2 negative and 1 positive delays m_k in the exponent, since such terms are the only ones that occur in the interesting frequency region. The general expression includes all combinations with one more negative delay than positive. For example, for order 7 we must find all combinations of 3 positive and 4 negative frequency translations ω_0. There are, for each combination of delays,

$$n_p = \binom{p}{\lfloor p/2 \rfloor}$$

combinations of x_1 and x_2 to consider. For example, for $p = 5$ we have $n_5 = 10$ combinations as shown above, for $p = 7$ we have $n_7 = 35$ combinations, for $p = 9$ we have $n_9 = 126$ combinations, and so on. Each combination consists of the correct number of x_1 and x_2 factors, where each x_2 factor must have a corresponding complex-conjugate x_2^*, so that there are always an even number of x_2-based factors in the expression, and an odd number of x_1. The overall product is rotated by a complex constant that depends on the time delays, and all terms of a given order *p* are summed.

The complete series: The complete expression consists of the weighted sum of all terms of all orders and delays, see the equation on the top of the page (up to order 5). For completeness, we have published MATLAB m-files with the correct representation up to an order of 9 [14].

3.1 Sampling frequency

An important discussion is which sampling frequency that will be needed for the DPDs. The original expression (1) has a sampling frequency that must be at least twice the bandwidth of the dual-band signal, including all bandwidth expansion due to the nonlinear amplifier, which may be a very large bandwidth indeed. However, if we study the expression of the dual-input predistorters we see that the terms are products of up to *p* narrow-band signals, and that we therefore can downsample the output signal y_n to *p* times the narrowband signal bandwidth, *W*. This is a huge reduction compared to the original sampling requirements, and this is the entire reason for the proposed dual-band linearizer.

4. COMPLEXITY-REDUCED FORMS

Since the full dual-band Volterra above can be too complex for practical use, we here develop a few complexity-reduced versions based on reducing the set of cross-terms in the series.

4.1 Complex-conjugate terms

First, we will enforce all complex-conjugate terms to have the same delay as the non-complex-conjugates. By this operation, there will always be pairs of $x_{n-m}x_{n-m}^* = |x_{n-m}|^2$, thereby allowing real-valued multiplications instead of complex ones, and fewer delays to sum over. We also combine the constants $a_{m_1 \dots m_p}^{(p)}$ and the complex rotations in the general form into new complex constants $b_{m_1}, b_{m_1 m_2}^{(s_1)} \dots$, see(2). By this combination we will not exploit the full structure of the problem since the new constants will now be correlated if chosen correctly. From a complexity-perspective the lost structure is of small help anyway, so the nice form of the new equation well motivates this simplification. The new complexity-reduced form is given by:

$$\begin{aligned} y_n &= \sum_{m_1=0}^{M} b_{m_1} x_{1,n-m_1} \\ &+ \sum_{m_1=0}^{M} \sum_{m_2=0}^{M} \sum_{s_1=1}^{2} b_{m_1 m_2}^{(s_1)} x_{1,n-m_1} \left| x_{s_1,n-m_2} \right|^2 \\ &+ \sum_{m_1=0}^{M} \sum_{m_2=0}^{M} \sum_{m_3=0}^{M} \sum_{s_1=1}^{2} \sum_{s_2=s_1}^{2} b_{m_1 m_2}^{(s_1,s_2)} x_{1,n-m_1} \left| x_{s_1,n-m_2} \right|^2 \left| x_{s_2,n-m_3} \right|^2 \end{aligned} \quad (2)$$

The generalization to higher orders than 5 can be easily understood from this equation.

4.2 Remove cross-signal cross-terms

By further restricting the allowed cross-terms between different delays and signals, we arrive at an even simpler form,

$$\begin{aligned} y_n &= \sum_{m=0}^{M} \sum_{k=0,2,\dots}^{P} \sum_{j=0,2,\dots}^{k} b_{m,0}^{(k,j)} x_{1,n-m} \left| x_{1,n-m} \right|^{k-j} \left| x_{2,n-m} \right|^{j} \\ &+ \sum_{m=0}^{M} \sum_{l=1}^{G} \sum_{k=2,4\dots}^{P} \sum_{j=0,2,\dots}^{k} b_{m,l}^{(k,j)} x_{1,n-m} \left| x_{1,n-m-l} \right|^{k-j} \left| x_{2,n-m-l} \right|^{j} \\ &+ \sum_{m=0}^{M} \sum_{l=1}^{G} \sum_{k=2,4\dots}^{P} \sum_{j=0,2,\dots}^{k} b_{m,-l}^{(k,j)} x_{1,n-m-l} \left| x_{1,n-m} \right|^{k-j} \left| x_{2,n-m} \right|^{j} \end{aligned}$$

which is the concurrent dual-band form of the generalized memory polynomial (GMP) [15]. We denote this form the dual-band-GMP (DB-GMP). In the traditional GMP description, the equation is often extended with non-bandlimited terms[1] by allowing for odd values of *k* and *j*, which can improve the performance in some applications.

[1] Terms that contain magnitude-squared factors are bandlimited to *p* times the bandwidth of the input signal, since we can write such terms by *p* multiplications. Terms that contain the magnitude raised

4.3 Remove all cross-terms

Removing the cross-terms altogether leads to the dual-input memory polynomial,

$$y_n = \sum_{m=0}^{M} \sum_{k=0,\ldots}^{P} \sum_{j=0,2,..}^{k} b_m^{(k,j)} x_{1,n-m} \left| x_{1,n-m} \right|^{k-j} \left| x_{2,n-m} \right|^{j}$$

This is the form proposed by Bassam et al in [4], and when used for predistortion it is usually denoted 2D-DPD (in the proposal by Bassam, also non-bandlimited terms are included, by allowing for odd *k* and *j*).

4.4 Remove all memory

Finally, by removing all memory we arrive at the dual-input polynomial (DIP),

$$y_n = \sum_{k=0,\ldots}^{P} \sum_{j=0,2,..}^{k} b^{(k,j)} x_{1,n} \left| x_{1,n} \right|^{k-j} \left| x_{2,n} \right|^{j}$$

4.5 Other complexity-reductions

There are of course also other complexity-reductions of the full Volterra that can be considered, such as:

- If the x_2 signal has much less power than x_1 (or vice versa), it may be useful to only include the linear effects of x_2, since higher order terms may be insignificant.
- The memory depth of high-order terms can be reduced compared to the linear terms, since high-order, high-delayed terms should be of less significance.
- Likewise, the nonlinear order of high-delayed terms can be reduced.
- In general, the general dual-input Volterra can be pruned by excluding insignificant terms in various ways.

We leave studies of other complexity-reduced forms to future work.

5. SIMULATIONS

We have performed simulations using a model derived from a commercial Mini-Circuits amplifier, and compared the results when different techniques for the linearization are used. In Fig. 2, we illustrate the spectrum, before linearization and linearized with the proposed DB-GMP technique. It is clear that the spectral regrowth is

In the table below, we show normalized mean square error (NMSE) results for various techniques. We see that the traditional single-input DPD fails to linearize, but that the 2 dual-input techniques both work fine. Among the dual-input techniques, the DB-GMP technique has a small advantage.

to an odd number are not bandlimited, since a square-root operation is necessary.

Technique	NMSE
No DPD	-17.6 dB
Single-input DPD	-25.0 dB
2D-DPD	-46.6 dB
DB-GMP	-47.6 dB

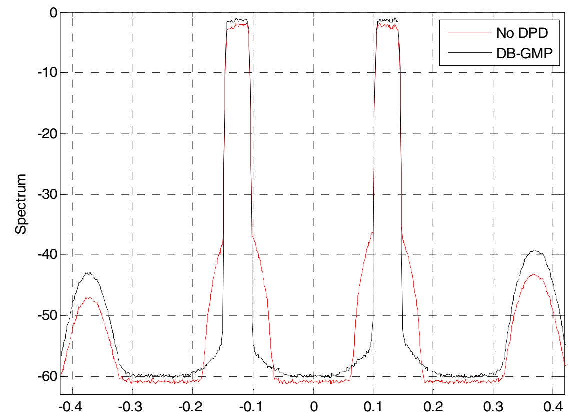

Fig.2. The output spectrum before (red) and after (black) linearization. The DB-GMP algorithm can effectively cancel the spectral regrowth around both the narrowband signals.

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7. REFERENCES

[1] P. Colantonio, *et al.*, "A design technique for concurrent dual-band harmonic tuned power amplifier," *Microwave Theory and Techniques, IEEE Transactions on,* vol. 56, pp. 2545-2555, 2008.

[2] P. Saad, *et al.*, "Design of a concurrent dual-band 1.8–2.4-GHz GaN-HEMT Doherty power amplifier," *Microwave Theory and Techniques, IEEE Transactions on,* vol. 60, pp. 1840-1849, 2012.

[3] S. Bassam, "Advanced signal processing techniques for impairments compensation and linearization of SISO and MIMO transmitters," *Doctor Dissertation,* 2010.

[4] S. A. Bassam, *et al.*, "2-D digital predistortion (2-D-DPD) architecture for concurrent dual-band transmitters," *Microwave Theory and Techniques, IEEE Transactions on,* vol. 59, pp. 2547-2553, 2011.

[5] S. A. Bassam, *et al.*, "Linearization of concurrent dual-band power amplifier based on 2D-DPD technique," *Microwave and Wireless Components Letters, IEEE,* vol. 21, pp. 685-687, 2011.

[6] H. Ku and J. S. Kenney, "Behavioral modeling of nonlinear {RF} power amplifiers considering memory effects," *IEEE Transactions on Microwave Theory and Techniques,* vol. 51, pp. 2495-2504, December 2003 2003.

[7] Y.-J. Liu, *et al.*, "Low-complexity 2D behavioural model for concurrent dual-band power amplifiers," *Electronics Letters,* vol. 48, pp. 620-621, 2012.

[8] Y.-J. Liu, *et al.*, "Digital Predistortion for Concurrent Dual-Band Transmitters Using 2-D Modified Memory Polynomials," 2013.

[9] H. Xiang, *et al.*, "Dynamic deviation reduction-based concurrent dual-band digital predistortion," *International Journal of RF and Microwave Computer-Aided Engineering,* 2013.

[10] D. Mirri, *et al.*, "A nonlinear dynamic model for performance analysis of large-signal amplifiers in communication systems," *IEEE Transactions on Instrumentation and Measurement,* vol. 53, pp. 341-350, 2004.

[11] A. K. Kwan, *et al.*, "Concurrent dual band digital predistortion using look up tables with variable depths," in *Power Amplifiers for Wireless and Radio Applications (PAWR), 2013 IEEE Topical Conference on*, 2013, pp. 25-27.

[12] S. Zhang, *et al.*, "A Time Misalignment Tolerant 2D-Memory Polynomials Predistorter for Concurrent Dual-Band Power Amplifiers," 2013.

[13] H. Cao, *et al.*, "Digital Predistortion for High Efficiency Power Amplifier Architectures Using a Dual-input Modeling Approach," *IEEE Transactions on Microwave Theory and Techniques,* vol. 60, pp. 361-369, 2012.

[14] *Modeling of Power Amplifiers*. Available: thomaseriksson-software.wikispaces.com/GHz+Centre

[15] D. R. Morgan, *et al.*, "A Generalized Memory Polynomial Model for Digital Predistortion of {RF} Power Amplifiers," *IEEE Transactions on Signal Processing,* vol. 54, pp. 3852-3860, 2006.