2004

Predistorter using overcompensation for power amplifiers

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DOI: <https://doi.org/10.31979/etd.m7st-j9jh>
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PREDISTORTER USING OVERCOMPENSATION
FOR
POWER AMPLIFIERS

A Thesis

Presented to

The Faculty of the Department of Electrical Engineering

San Jose State University

In Partial Fulfillment

Of the Requirements for the Degree

Master of Science

By

Ralph Inducta

May 2004

UMI Number: 1420431

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ABSTRACT

PREDISTORTER USING OVERCOMPENSATION FOR POWER AMPLIFIERS

By Ralph Inducta

Performance of a simulated, broadband linearization technique at 5 GHz using a modified form of feedforward is presented. The technique utilizes the driver amplifier as a generator of nonlinear distortion to compensate for the distortion components of the nonlinear device such as the amplifier. Comparisons were made, and insights were given between the two linearization techniques. 20 MHz bandwidth at upper U-NII band was used, and the modified form of feedforward achieved 22 dB suppression. Other bandwidths in 5 GHz range were tested, and achieved a modest 14 dB suppression.

Acknowledgements

I would like to thank my mentor, Dr. Somnath Mukherjee, for lending his never-ending ideas. His guidance and support is always there for me to count on. For that, I'm truly blessed and grateful.

I would also like to thank Prof. George Vendelin for everything that an advisor can provide and beyond. Thank you for supporting my ideas.

My gratitude goes to Prof. Robert Morelos-Zaragoza for his eagerness to help on a very short notice.

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Introduction:

Feedforward technique is widely used for improving the linearity of amplifiers. Advantages of using feedforward include broadband and higher distortion suppression. Feedforward application began in the work of Seidel et al.¹ Since then, there are numerous articles analyzing feedforward type linearizers.²⁻⁴

Predistortion technique is also used for linearization to reduce the circuit output's ACP value.⁵ Predistortion techniques are usually simple to implement, but achieve modest suppression of distortion.⁶ One article provides striking resemblance to the proposed technique where the driver amplifier generates IMD distortion to compensate the device under test.⁷ The article showed result of 9 dB suppression of spectral regrowth at 2.37-2.4 GHz band.

The proposed technique is a deviation from a generalized form of feedforward, and operates as a predistorter. Therefore, the technique is applicable to systems where the input is a different form than the output. Examples include application in up conversion stage and broadband linearization of electro-optic device. The following advantages of the technique include:

- 1) Losses due to the delay line and couplers in the main path are eliminated.
- 2) Low cost delay line can be used.
- 3) Efficiency is higher compared to FF technique.

There is tremendous work done in linearization at 2.4 GHz band, and future work is growing at 5 GHz band where there is no congestion and stiff competition. Already, 5 GHz is the platform band of WLAN (802.11a), U-NII band in the USA, and HIPERLAN/ISM in Europe. There is also a need for linearization for phase-array antenna applications.⁸ An example of the proposed technique is simulated at 5 GHz system.

I. Amplifier Distortion

Nonlinear elements such as amplifiers can be modeled using Taylor Polynomial series. Power series is used as a mathematical analysis for its simplicity, but requires a simplifying assumption that the circuit contains only ideal memoryless transfer nonlinearities.⁹ The output voltage transfer function $V_{out}(t)$ is the sum of all N^{th} order input voltage components of an element.

$$V_{out}(t) = \sum_{n=1}^N C_n V_{in}^n(t) \quad \text{Eqn. I.1}$$

where C_n are the Taylor Polynomial coefficients, and $V_{in}^n(t)$ are the input voltages of the particular harmonics.

Theoretically, the summation of the harmonic voltages goes to infinity. For simple illustration, the summation of the harmonic voltages is terminated at some point N .

Equation I.1 shows an amplifier generates both even and odd order harmonics. As an example for $N = 5$, equation I.1 gives the output voltage $V_{out}(t)$ as:

$$V_{out}(t) = \sum_{n=1}^5 C_n V_{in}^n(t) = C_1 V_{in}(t) + C_2 V_{in}^2(t) + C_3 V_{in}^3(t) + C_4 V_{in}^4(t) + C_5 V_{in}^5(t)$$

$$\text{Eqn. I.2}$$

Coefficient C_1 represents the gain factor of the device, while the 2nd and 4th order represents the even terms. The first term falls on the fundamental frequency f , the second term lands on frequency $2f$, third term lands on frequency $3f$, and so on, as illustrated in Figure 1.

Since for Class-A amplifiers, the dominant orders are the 2nd and 3rd, equation I.2 becomes

$$V_{out}(t) = C_1 V_{in}(t) + C_2 V_{in}^2(t) + C_3 V_{in}^3(t) \quad \text{Eqn. I.3}$$

The first element in equation I.3 is the input voltage scaled by a factor C_1 , while the 2nd and 3rd elements are the 2nd and 3rd harmonic distortion respectively. As an example, signal $V_{in}(t) = 2 \cos(\omega t)$ with frequency 1.5 GHz is passed through an amplifier. Amplifier generates infinite harmonics of the input signal $V_{in}(t)$ at $2f = 3$ GHz, $3f = 4.5$ GHz, and so on. Figure 2 illustrates this example, given $C_1 = 10$, $C_2 = 5$, $C_3 = 3$, $C_4 = 2$, $C_5 = 4$. Figure 2 shows harmonics up to 5th order. Notice the amplitude of 3rd harmonics is higher compared to higher N^{th} -order harmonics. Notice also from Figure 2, the output signal $V_{out}(t)$ is distorted, leading to the generation of harmonic distortions.

I.A 2-Tone Test

2-tone test has been used in the RF industry to determine the distortion performance of a device. When two signals with equal amplitudes but different frequencies are applied to a nonlinear device such as an amplifier, beating tones interact to produce intermodulation distortions. Intermodulation distortions, or IMD, are unwanted signals which cause interference in the bandwidth of interest. Combination of these two signals will arise into mixing frequencies $f_{m,n}$, where m and n are the orders of the mixing products and f_1 and f_2 are the applied frequencies.

$$f_{m,n} = m \cdot f_1 \pm n \cdot f_2 \quad \text{Eqn. I.4}$$

For example, 2nd order IMD would generate frequencies at:

$$\begin{array}{l} 2f_1 \\ 2f_2 \\ f_1 \pm f_2 \end{array} \quad \text{Eqn. I.5}$$

For a 3rd order IMD, generated frequencies are:

$$\begin{array}{l} 3f_1 \\ 3f_2 \\ 2f_1 \pm f_2 \\ 2f_2 \pm f_1 \end{array} \quad \text{Eqn. I.6}$$

Figure 3 shows the 3rd order IMD products falling at $2f_1-f_2$ and $2f_2-f_1$ frequencies. A similar procedure can then be applied for higher order IMD, such as the 5th order.

I.B Intercept Points

For a nonlinear device such as the amplifier, linear output power is a replica of the input power with just a gain factor of C_1 as depicted in equation I.1. 2nd order IMD or IM2 increases 2x as fast as the fundamental power (IM1). At a certain point, both the linear output and IM2 intersect. This intersection is called *2nd order Intercept Point*, or IP2. Similarly, 3rd order IMD or IM3 increases 3x as fast as the fundamental power, and intersects at the *3rd order Intercept Point*, or IP3. Intercept Points are referred either at the input or the output. Front-end applications such as High Power Amplifier (HPA) have their IP's referred at the output, referred as *OIP2* and *OIP3*. In reality, this point can't be reached. There would come a point where fundamental power starts to deviate from its linear path, and saturates. DC power supply can't supply enough power to the device for its fundamental power to reach the intercept

points. Thus, output power saturates. When a 1dB/1dB change is no longer true, a point called *1-dB Compression Point*, or P_{1dBc} has been reached.

Figure 4 illustrates the input-output power relationship of an amplifier. IM1 is the fundamental power that increases linearly with the input power before P_{1dBc} . P_{sat} refers to the saturation gain of the amplifier, as shown in Figure 4. IM2 increases at the rate 2x as that of the fundamental power. Accordingly, IM3 increases at the rate 3x as fast as the fundamental power. Intercept Points (IP's) are figure of merit to determine the level of IM distortions at a fixed output or input power levels. Equation I.7 shows the relationship of OIP to the IM output power and the output power of the signal.

$$P_{out_{IMn}} = n \cdot P_{OUT} - (n - 1) \cdot OIP_n \quad \text{Eqn. I.7}$$

The term n is the N^{th} -order, $P_{out_{IMn}}$ is the IM output power of the N^{th} -order, and OIP_n is the output intercept point of the N^{th} order. For example, if the value of the IM output power is desired for 3rd order IMD, equation I.7 becomes:

$$P_{out_{IM3}} = 3 \cdot P_{out} - 2 \cdot OIP_3 \quad \text{Eqn. I.8}$$

II. Theory of Operation

Feedforward technique has been widely used particularly in amplifiers for its excellent suppression of intermodulation distortions over a wide bandwidth. The general operation in feedforward technique is to suppress the main carrier signals in the first loop, amplify the distortion created by amplifier A1 in Path 4 with opposite phase, and recombine the signals at C3 output as shown in Figure 5. If the distortion vectors in Path

3 and Path 4 are equal in magnitude but opposite in phase, then cancellation of distortion occurs.

Although feedforward has been widely accepted, there are several disadvantages of using feedforward technique. This technique requires low insertion loss on the main path after amplifier A1 to achieve higher efficiency. Having lower insertion loss especially on the delay line is costly. The proposed technique solves that problem.

The modified technique has the same 1st loop operation as the feedforward. The general block diagram of this technique is shown in Figure 6. Path 1 and Path 2 both have the main carrier signals, equal in magnitude but opposite in phase at H1 output. The difference between techniques is in the 2nd loop, where amplifier A2 has higher gain compared to feedforward. The IM signals generated at point A is then amplified more than that of feedforward in Path 4, as shown in Figure 5. IM vector of amplifier A3 in Path 3 is added with opposite IM vector of amplifier A2 in Path 4, leaving a resultant distortion vector, as illustrated in Figure 7 (a). At point C, the A1-generated IM vector is added with this residual IM vector distortion. The addition of vectors leave a small distortion compared to A1-generated distortion, as shown in Figure 7 (b). The residual distortion at point B is due to overcompensating the feedforward, which is appropriately term *OverCompensating Feedforward* (OCFF), and act as a predistorter.¹⁰

III. Passive Components

III.A Directional Couplers

Directional couplers can be implemented in a couple of ways. For frequencies less than 1 GHz, directional couplers can be realized using transformers. For frequencies higher than 1 GHz, microstrips or striplines are used.

For 5 GHz system, a quadrature stripline coupler had been designed. As an example, a single section, coupled line stripline coupler design as described by Pozar¹¹ was realized with the substrate dielectric constant of 3.48. Dielectric constant of 3.48 is typical for a *Rogers 4000 series* substrate. An example design is simulated in Microwave Office (MWO).

EM design was created using actual layout as shown in Figure 8 (a). Three dimensional layout of the design is shown in Figure 8 (b). Edge ports with 1 mm lengths were added for each of the 4 ports. Dimensions are:

Thickness: 0.15 cm Length: 0.9 cm

Width: 0.182 cm Separation: 0.00627 cm

Figure 9 (a) illustrates a coupling loss of 10 dB (square), and at the same time, introduces almost a dB of insertion loss (triangle). Phase introduces almost 90 degrees between the output port and the coupling port, as shown in Figure 9 (b).

III.B. Delay Lines

Delay lines are important in designing linearization techniques such as feedforward. Delay lines are used to compensate the electrical delays caused by each component in the other path of the signal. There are various ways to implement variable delay lines.¹²

Besides from linearization applications, delay lines are used in variety of applications, such as phased-array antennas, ECM systems, and digital applications. Size, complexity, cost, and insertion loss are the parameters to consider when choosing delay lines.

Because of its sizes, digital delay lines are used in feedforward applications. Digital delay lines consist of delay lines of various lengths to provide various electrical delays, connected to switches. The switches provide the setting to which delays can vary. These delay lines can be coaxial cable, waveguide, striplines, or acoustical devices.

Waveguides are inappropriate because of their size and loss, and acoustical devices have matching issues according to Kirchner.¹²

Both the striplines and coaxial cable can be used to delay time signals in feedforward applications. Coaxial cables are preferable in higher time delays than striplines. Physical length of coax cable is easily adjusted, rather than changing the physical length of striplines.

For example, let the coax cable has the following characteristics:

Characteristic Impedance (Z_0) = 50 Ω

Dielectric Diameter (D)= 0.085 in

Dielectric constant (ϵ) = 2.03

Then the physical length L of the coax cable for the desired electrical delay T is¹³:

$$L = \frac{0.984 \cdot T}{\sqrt{\epsilon}} \quad \text{Eqn. III.2-1}$$

For the given coax cable characteristics, Figure 10 shows the relationship between the delay and length. For example, 12 ns delay requires 8 ft. of coaxial cable length.

IV. Variable Phase Shifters

Phase shifter can be implemented using a reflection-type phase modulator.¹⁴⁻¹⁵ Changing the voltages in the varactor diodes introduce phase changes. The use of hybrid coupler improves the varactor diode's return loss. A phase modulator using hybrid coupler based on alpha industries app note¹⁶ has been simulated. A SVM2019 nonlinear varactor diode model had been used instead of the suggested SVM1245 due to lower capacitance the SVM2019 provides compared to SVM1245. Lower varactor capacitance leads to higher dynamic phase range for this particular design. A nonlinear varactor model of SMV2019 from Alpha industries is given in Figure 11 (a).

Looking from Figure 11 (b), C_{tune} was added and optimized to decrease insertion loss and increase dynamic phase range of the variable phase shifter. Components L_{ser} , C_{par} , and C_{tune} are part of the matching circuit for the varactor diode's impedance. Applied reverse control voltage V_{ctl} was varied to introduce varying phase shifts. Hybrid device $HYB1$ simulates hybrid coupler device with gain and phase imbalances associated with such a device.

Figure 11 (c) shows various traces representing different values of control voltage V_{ctl} according to the Fig. 11 (b) design. Simulated result shows a good return loss with any values of control voltage V_{ctl} up to 12 V. Insertion loss is less than 2.5 dB on any given control voltages. From simulated result, adjusting the control voltage V_{ctl} affects the insertion loss of the device. Figure 11 (d) illustrates how control voltage affects the phase shift for a wide range of frequencies. For a particular frequency, 5.5 GHz for example, phase range is 50 degrees for this particular design. Phase range is expected to

change if further improvement on the matching circuit was done. The phase shifter design was also tested for linearity in phase. From Fig. 11 (e), deviation from linear phase DLP shows within ± 15 degrees deviation across the voltage range of V_{ct1} . Therefore, this design is ideal for applications where linearity in phase like the linearization technique is a requirement.

If more than 180 degrees of phase change is required, a cascade of this device would produce such result, as shown in Figure 12 (a). SVM2019 voltages are varied to produce desired phase changes. For example, if phase shift of 248 degrees is required, a control voltage V_{ct2} (2nd control voltage) is set to 1 V for 5.5 GHz frequency, as shown in Figure 12 (b). Figure 12 (b) illustrates the phase range with different frequencies, at control voltage $V_{ct1} = 1$ V.

An ideal phase shifter model from HP ADS with the same phase shift was compared to the cascaded phase shifter design. Figure 12 (c) shows the same phase change of the phase shifter design with the ideal phase shifter at 5.5 GHz.

By changing C_{par} to 1pF on the second phase shifter design, phase shift from -129 degrees to about -118 degrees can be achieved when $V_{ct1} = 0.2$ V is set, as shown in Figure 12 (d). Variations of V_{ct1} , V_{ct2} , and the matching circuitry will provide the desired phase range from the cascaded phase shifter design.

V. 1st Loop

In feedforward schemes, the 1st loop's main function is to cancel the input power carrier signals. Overcompensated scheme functions the same way. Time delay caused

by Path 1, mostly coming from amplifier A3, should set equal to delay line *DL1* time delay, according to Eqn. A-8. Another requirement according to Appendix A for suppressing carrier signals is Att1 be set equal to all the losses and gain from both Path 1 and Path 2, as shown in Figure 13.

In simulated 5 GHz system, ADS system amp models were used to simulate the nonlinear properties of the amplifiers, assuming all matching circuits had been designed. Coupler design with performance from Figure 18 (a) – (b) have been used.

Assumption of 9 ns group delay of A3 was simulated by the time delay *A3_GD* element if all matching circuits of A3 were designed. Delay line *DL1* was simulated using a time delay assumption of $\text{delay1} = 9 \text{ ns}$, and *DL1_loss* assumption of 3 dB.

Using 2-tone test, simulated result showed a 19 dB suppression of main carrier signals at 5.5 and 5.6 GHz, as shown in Figure 14. Input power was set to 20 dBm. Main signals are the blue trace (marker m3), while the red trace represents the closed loop output power at the hybrid *HYB1* output (marker m1). Around 19 dB of carrier suppression was achieved with this design in 5-6 GHz range.

VI. 2nd Loop

The 2nd loop's main function is to cancel the distortion generated by A3 amplifier at the output of the system in FF system. Referring to Fig. 5, distortion generated by amplifier A1 in Path 3 is combined with distortion in Path 4. Delay in Path 4 should be equal to delay line in Path 3, and attenuator ATT2 is set equal according to the losses as described in Appendix B.

In OCFF system, A2 requires higher gain than that of feedforward to overcompensate the distortion at point B, referring to Figure 15.

$$\text{Let } \text{OIP3}_{A3} = A3 \text{ OIP3} \quad \text{OIP3}_{A1} = A1 \text{ OIP3}$$

$$L = \text{Path 3 insertion loss} \quad G_{\text{shunt}} = \text{extra gain minus losses in Path 4}$$

$$lc2, lc3 = \text{insertion losses of couplers C2 and C3}$$

where

$$L(\text{dB}) = 10 \cdot \log \left(\frac{1}{lc2 \cdot DL2_{\text{loss}} \cdot lc3} \right) \quad \text{Equation VI-1}$$

$$G_{\text{shunt}}(\text{dB}) = 10 \cdot \log \left(\frac{g_{A2}}{c_3 \cdot Att_2 \cdot c_2 \cdot h_1} \right) \quad \text{Equation VI-2}$$

In order to over-compensate the distortion at point B, G_{shunt} would clearly be higher than the losses in Path 3. Thus, $G_{\text{shunt}}(\text{dB}) \neq L(\text{dB})$. As explained by Mukherjee et. al¹⁰, in terms of amplifiers' IP3's and the relation to G_{shunt} , G_{shunt} therefore equal to

$$G_{\text{shunt}}(\text{dB}) = L + 20 \cdot \log \left(1 + 10^{\frac{(\text{IP3}_{A3} + L) - (\text{IP3}_{A1} - G_{A1})}{10}} \right) \quad \text{Equation VI-3}$$

For $L = 7 \text{ dB}$, and $A1 = 10 \text{ dB}$ gain, Figure 16 shows the relationship of the extra gain needed with respect to A3 TOI to compensate for the distortion introduced by A1.

Deducing from the graph, as A3 TOI increases, the higher the extra gain (ie. G_{shunt}) is needed because the IM distortion produced by A3 decreases. Also, for a specific A3

TOI, the lower A1 OIP3 is, the higher extra gain is needed because the IM distortion generated by A1 is increasing.

VII. 5 GHz OCFF Performance

The 5 GHz OCFF system performance is described. As an example, 2 tone input frequencies at $f_1 = 5.765$ GHz and $f_2 = 5.785$ GHz were used, which falls under the upper band, U-NII WLAN bandwidth. Amplifiers in this system are system models from HP ADS. A device under test is a 1.26W device, and +45 dBm TOI with 9 ns estimated group delay, shown in Figure 17 (a). Amplifier A1 model doesn't contain group delay, therefore a time delay A1_GD element was introduced. The system model doesn't also have biasing options.

2-tone test were applied with 20 MHz separation. The device shown in Figure 17 (a) was operated as Class-A with 10 dB gain, and third order intermodulation distortions (IM3) were observed at $2f_1 - f_2$ and $2f_2 - f_1$. Output spectrum showed $P_{out} = 28$ dBm with third-order intermodulation (TOI) of +43.361 dBm and $IM3 = 30.148$ dBc. The term dBc is used to determine the level of distortion relative to the carrier signal.

In 5 GHz simulated system, a coupler was designed to have the performance shown in Figure 18 (a). The coupler has ~ 0.5 dB insertion loss, good return loss, and 10 dB coupling. 90 degrees phase difference between the output port and the coupling port is also shown. From Figure 18 (b), the coupler also introduces 43 ps delay.

The coupler was used throughout the 5 GHz simulated system, designated as C1, C2, and C3. Phase Shifter PS2 was also used and described in previous section. Delay lines

DL1 and *DL2* were modeled using a pad loss with time delay elements. Group Delay *A1_GD*, *Error_Amps_GD*, and *A3_GD* are amplifiers' group delay followed by their respective amplifier models. Group Delay *Error_Amps_GD* is the sum of all the group delays introduced by amplifiers' *A2_1*, *A2_2*, *A2_3*, and *A2_4*. Amplifier *A3* can be thought of as a driver for amplifier *A1*. HYB1 is an ideal 180 degrees hybrid with amplitude imbalance of 0.5 dB, 5 degrees phase imbalance, and introduces 60 ps delay. All of the error amplifiers and the DUT nonlinear parameters are close to real-life components' parameters as possible, assuming proper matching circuitry and biasing were applied. The error amplifiers can be linear, small-signal amplifiers because it only amplifies IM distortion, assuming good signal suppression from 1st loop. Attenuator *ATT1* was adjusted to get optimized carrier suppression in the first loop, while attenuator *ATT2* was adjusted for good IM suppression. Input power *Pin* is set to 20 dBm per carrier. The completed design is shown in Figure 19.

VII.A A3 TOI

OCFF theory of operation requires higher gain in Path 4 than necessary in FF with respect to A3 TOI as described previously. In 5 GHz simulated system, this theory is explored with good IM suppression in mind. Theoretically, as A3 IM distortion decreases, the higher extra gain from the A2 cascade amplifiers is required. A3 TOI of +41.75 dBm was chosen, and according to Figure 16, the shunt path gain (G_{shunt}) in Path 4 would need +12.8 dB gain for A1 OIP3 of +45 dBm. Error amplifiers were chosen such that there is enough gain in Path 4 to overcome Path 3 losses, and still have enough dynamic range for the attenuator settings.

At this particular A3 TOI value with Att1 = 10 dB, Att2 = 17 dB, and Path 3 loss = 7 dB, the simulated result needed +12.5 dB gain to have good IM3 suppression. The simulated G_{shunt} gain is very close to the theoretical value. If A3 TOI value is increased, the expected simulated G_{shunt} will follow the theoretical graph, as shown in Figure 20.

VII.B IM Relationship between FF and OCFF

As previously explained, higher gain in Path 4 is needed to have extra amount of distortion at the C3 output. Therefore, higher gain leads to higher IM distortion levels. If the extra distortion is equal in magnitude but opposite in phase to that of amplifier A1 distortion, then suppression of A1 amplifier is achieved. This is best illustrated with the graph in Figure 21 from 5 GHz simulated system. “OCFF” data is taken at the output of A1, while the “FF” data is taken at the output of C3 with A3 amplifier as the DUT as in feedforward case. Phase Shifter *PS2* was maintained at -123 degrees, while attenuator ATT2 stayed at 20 dB while output power was swept. Setting at ATT2 = 20 dB provided better distortion performance at C3 output as expected in FF case. However, this setting is not optimum for feedforward case. Lower extra gain in Path 4 in FF case compared to OCFF case is as expected. Notice the increase in IM distortion in OCFF case while the FF case had a lower IM distortion.

The FF case had a non-monotonic graph between $P_{\text{out}} = 26$ dBm and 29 dBm. This region was optimized to have good suppression at C3 output. The region after $P_{\text{out}} = 30$ dBm increases because saturation point for A1 had been reached. AM-AM and AM-PM occurs in this region.

The IM relationship between FF and OCFF is clearly illustrated from Figure 22. Notice the big jump in IM distortion in OCFF when gain is decreased. On the other hand, notice also the IM distortion in FF output lowered when gain is also decreased. Thus, 5 GHz simulated system follows the OCFF theoretical approach given previously.

VII.C IM Suppression

After the system was optimized, with $A1 \text{ TOI} = +41.75 \text{ dBm}$, the 5 GHz system achieved higher TOI with the worst case of 22 dB IM3 suppression. TOI value increased by 11 dB, as shown in Fig. 23 (a). Notice the output power back-off is 3 dB, since $A1 P_{\text{sat}}$ is 31 dBm.

The suppression performance of OCFF is comparable to that of feedforward system. The IM3 at frequency $2f_1 - f_2$ can be suppressed more, but at the expense of degrading the IM3 suppression at frequency $2f_2 - f_1$. The settings were chosen to balance IM3 for both frequencies. Markers $m21$ and $m22$ are IM3 distortions from the DUT, while markers $m2$ and $m23$ are the OCFF system distortions. $\text{TOI}_{\text{rightside}}$ and $\text{IM3}_{\text{rightside}}$ refers to frequency $2f_2 - f_1 = 5.805 \text{ GHz}$. Attenuator settings for ATT1 is 10 dB, and $\text{ATT2} = 17.35 \text{ dB}$. Control voltages for phase shifter V_{ct1} and V_{ct2} is 0.55V and 0.25 V respectively. The control voltage settings correspond to phase shift of -123 degrees.

The simulated system is not only applicable for 20 MHz bandwidth. Other frequencies with different bandwidth between 5-6 GHz had been simulated. For example, $f_1 = 5.5 \text{ GHz}$ and $f_2 = 5.7 \text{ GHz}$ was used for 2-tone test with a worst case result of 15 dB IM suppression, shown in Figure 23 (b). Settings are: $\text{Att2} = 17.35 \text{ dB}$, $V_{\text{ct1}} = 0.2 \text{ V}$, $V_{\text{ct2}} = 0.55 \text{ V}$. These control voltage settings correspond to phase shift of around

-125 degrees. For 100 MHz bandwidth at the lower end of 5 GHz spectrum, 14 dB suppression is achievable. Two frequencies at $f_1 = 5.2$ GHz and $f_2 = 5.3$ GHz were applied and tested, shown in Figure 23 (c). Settings for this test are: $Att2 = 16.35$ dB, $V_{ct1} = 0.25$ V, and $V_{ct2} = 0.55$ V.

VII.D Noise

Assuming 2nd loop is perfectly balanced, the noise introduced by amplifier A3 would cancel at C3 output because the added noise in Path 3 would also be present in Path 4 but opposite in phase. Therefore, the noise added into the system flows from Path 2 into Path 4, and goes out of C3 output. In 5 GHz simulated system, error amplifiers were arranged in such a way that the amplifier chain introduces low noise figure and higher TOI. The total expected noise figure of the error amplifier chain is 1.94 dB. At the same time, error amplifier chain provides TOI of +40 dBm.

VIII. Efficiency

One of the advantages of using OCFF technique is the higher efficiency it provides over FF. In order to compare both systems, it is appropriate to assign DC values. The amplifiers' DC biasing was determined to match as close as possible to that of real components. Assumptions were made for comparison purposes. The following table lists the DC power supply values.

A1 =	9 Vdc		A2_1 =	4 Vdc		A2_3 =	5 Vdc
	323 mA			75 mA			175 mA
A3 =	9 Vdc		A2_2 =	5 Vdc		A2_4 =	9 Vdc
	270 mA			65 mA			130 mA

Table 1 DC Power Supplies

The efficiency for both systems is very low because all amplifiers are in Class-A operations. In this case, linearity is the main function of the linearization technique. But for comparison purposes, the following is illustrated.

Let P_{A1} = A1 output power

I_m = losses in Path 3

P_{dcA1} and P_{dcA2} are the dc power consumptions of A1 and A2 respectively.

Then feedforward efficiency is:

$$\eta_{FF} = \frac{\frac{P_{A1}}{I_m}}{P_{dcA1} + \sum P_{dcA2}} \quad \text{Eqn. VIII-1}$$

For OCFF system, overall efficiency where P_{dcA3} is A3 dc power consumption is:

$$\eta_{OCFF} = \frac{P_{A1}}{P_{dcA1} + \sum P_{dcA2} + P_{dcA3}} \quad \text{Eqn. VIII-2}$$

From Figure 24, OCFF efficiency is greater than the FF efficiency after 2 dB loss in the main path. OCFF efficiency is lower before 1.5 dB loss in the main path due to the contribution of DC power consumption from amplifier A3. Notice in order to have feedforward efficiency higher than OCFF, main path loss (Path 3 loss) would have to be

less than 1.5 dB. Assuming the insertion losses of C2 and C3 combined amount to 1 dB, then the main delay line only have a maximum of 0.5 dB loss at 5 GHz. Delay line of 0.5 dB translates into expensive real estate. With OCFF, cheap, lossy delay line can be used and still achieve higher efficiency compared to feedforward, as described in delay line section.

OCFF efficiency is not a function of the main path loss because signals are applied at the input, not the output which degrades the A1 amplifier output power. Notice from the graph the feedforward dependency on the main path loss, particularly the delay line, to the overall efficiency.

Conclusion:

The 5 GHz OCFF system has been simulated and described. The study offered further insight on both feedforward and over-compensated feedforward technique. Phase shifter had been designed specifically for this simulation with good insertion loss and return loss. For U-NII upper band, 20 dB suppression is achievable. For other frequencies with different bandwidths, a modest 14 dB suppression is also achievable. The result from OCFF simulation will have an impact for system designers requiring cheaper delay lines and higher efficiency compared to that of feedforward at 5 GHz range.

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Appendix A

1st Loop Simplified Analysis

Simplified analysis of the general 1st loop block diagram is described as follows. All units are linear powers. Refer to Figure 13.

Let $lc1, lc2$ = insertion loss of coupler C1 and C2 C_1, C_2 = Coupling losses of couplers

$Att1$ = Attenuation of Attenuator h_1 = hybrid insertion loss

$DL1_{loss}$ = Delay line loss $DL1$ = Delay line

$V_{in}(t)$ = input voltage $V_{IM}(t)$ = IM distortion of A3

g_{A3} = gain of Amp A3 τ_{A3} = group delay of A3

$V_{error}(t)$ = closed loop voltage τ_{DL1} = electrical delay of DL1

At input port, point A, Let:

$$V_{in}(t) = A \cdot \cos(\omega_1 \cdot t) + A \cdot \cos(\omega_2 \cdot t), \quad A = 1 \quad \text{Eqn. A - 1}$$

where ω_1, ω_2 are the angular frequencies of frequencies f_1 and f_2 .

At output of A3:

$$V_{A3}(t) = \frac{g_{A3} \cdot e^{-j(\omega_1, \omega_2)\tau_{A3}}}{lc1} \cdot V_{in}(t) + V_{IM}(t) \quad \text{Eqn. A - 2}$$

Looking at Point D from Path 1:

$$V_{Path1} = \frac{1}{c_2 \cdot Att_1 \cdot h_1} \cdot V_{A3}(t) \quad \text{Eqn. A - 3}$$

Looking at Point D from Path 2:

$$V_{Path2} = \frac{-1}{c_1 \cdot DL1_{loss} \cdot h_1} \cdot V_{in}(t) \cdot e^{-j(\omega_1, \omega_2)\tau_{DL1}} \quad \text{Eqn. A - 4}$$

The negative sign is due to the reversal of phase caused by phase shifter β .

Combining the two paths together at point D:

$$V_{\text{error}}(t) = V_{\text{Path1}} + V_{\text{Path2}} \quad \text{Eqn. A - 5}$$

$$V_{\text{error}}(t) = \frac{1}{c_2 \cdot \text{Att}_1 \cdot h_1} \cdot \left[\frac{g_{A3} \cdot e^{-j(\omega_1, \omega_2)\tau_{A3}}}{|cl|} \cdot V_{\text{in}}(t) + V_{\text{IM}}(t) \right] - \frac{1}{c_1 \cdot \text{DLI}_{\text{loss}} \cdot h_1} \cdot V_{\text{in}}(t) \cdot e^{-j(\omega_1, \omega_2)\tau_{\text{DLI}}}$$

$$\text{Eqn. A - 6}$$

If the following equations are met,

$$\frac{g_{A3}}{c_2 \cdot \text{Att}_1 \cdot |cl|} = \frac{1}{c_1 \cdot \text{DLI}_{\text{loss}}} \quad \text{Eqn. A - 7}$$

$$\tau_{A3} = \tau_{\text{DLI}} \quad \text{Eqn. A - 8}$$

$$\text{then } V_{\text{error}}(t) = V_{\text{IM}}(t) \quad \text{Eqn. A - 9}$$

From the derived equations, all the losses and gain are usually fixed except for the attenuator Att1. Therefore, Att1 is set so that

$$\text{Att}_1 = \frac{c_1 \cdot \text{DLI}_{\text{loss}}}{c_2 \cdot |cl|} \cdot g_{A3} \quad \text{Eqn. A - 10}$$

Appendix B

2nd Loop Simplified Analysis

Simplified analysis for the 2nd loop block diagram is expressed in the following. All units are linear powers, and continued from the 1st loop voltage analysis from Appendix A. Refer to Fig. 15, assuming for a moment there is no amplifier A1 attached, as in FF scheme.

Let lc2, lc3 = Insertion loss of couplers C2, C3	C ₂ , C ₃ = Coupling loss of C2, C3
Att2 = Attenuation of Attenuator Att2	h ₁ = Insertion loss of Hybrid H1
DL2 _{loss} = Delay Line loss of DL2	DL2 = Delay line
g _{A2} = Gain of amp A2	g _{A1} = Gain of amp A1
τ _{A2} = Group Delay of A2	τ _{A3} = Group Delay of A3
V _{A3} (t) = output voltage of A3	V _{in} (t) = input voltage
V _{IM} (t) = IM voltage	V _{error} (t) = Error voltage at pt. D

At point A:

$$V_{A3}(t) = \frac{g_{A3} \cdot e^{-j(\omega_1, \omega_2)\tau_{A3}}}{lc1} \cdot V_{in}(t) + V_{IM}(t) \quad \text{Eqn. B - 1}$$

At point D, assuming perfect cancellation of main carrier signals:

$$V_{error}(t) = \frac{1}{c_2 \cdot h_1} \cdot V_{IM}(t) \quad \text{Eqn. B - 2}$$

Looking at point B from Path 3:

$$V_{\text{Path3}}(t) = \frac{1}{lc2 \cdot DL2_{\text{loss}} \cdot lc3} \cdot V_{A3}(t) \cdot e^{-j(\omega1, \omega2)\tau_{DL2}} \quad \text{Eqn. B - 3}$$

Looking at point B from Path 4:

$$V_{\text{Path4}}(t) = \frac{-1}{c_3 \cdot \text{Att}_2} \cdot g_{A2} \cdot e^{-j(\omega1, \omega2)\tau_{A2}} V_{\text{error}}(t) \quad \text{Eqn. B - 4}$$

Phase Shifter β reverses the sign in Path 4.

At point B:

$$V_{\text{outFF}}(t) = V_{\text{Path3}}(t) + V_{\text{Path4}}(t) \quad \text{Eqn. B - 5}$$

$$V_{\text{outFF}}(t) = \left[\frac{1}{lc2 \cdot DL2_{\text{loss}} \cdot lc3} \left(\frac{g_{A3} \cdot e^{-j(\omega1, \omega2)\tau_{A3}}}{lc1} \cdot V_{\text{in}}(t) + V_{\text{IM}}(t) \right) \right] \cdot e^{-j(\omega1, \omega2)\tau_{DL2}} - \left[\frac{1}{c_3 \cdot \text{Att}_2} \cdot \frac{1}{c_2 \cdot h_1} \cdot g_{A2} \cdot e^{-j(\omega1, \omega2)\tau_{A2}} \right] \cdot V_{\text{IM}}(t)$$

$$\text{Eqn. B - 6}$$

If the following equations are met as in feedforward case,

$$\frac{1}{lc2 \cdot DL2_{\text{loss}} \cdot lc3} = \frac{1}{c_3 \cdot \text{Att}_2 \cdot c_2 \cdot h_1} \cdot g_{A2} \quad \text{Eqn. B - 7}$$

$$\tau_{A2} = \tau_{DL2} \quad \text{Eqn. B - 8}$$

$$\text{then, } V_{\text{outFF}}(t) = \frac{g_{A3} \cdot e^{-j(\omega1, \omega2)(\tau_{A3} + \tau_{DL2})}}{lc2 \cdot DL2_{\text{loss}} \cdot lc3 \cdot lc1} \cdot V_{\text{in}}(t) \quad \text{Eqn. B - 9}$$

$$\text{if we let } g_f = \frac{g_{A3}}{lc2 \cdot DL2_{\text{loss}} \cdot lc3 \cdot lc1} \quad \text{Eqn. B - 10}$$

then $V_{\text{out}}(t) = (g_f \cdot e^{-j(\omega L/2)(\tau_{A3} + \tau_{DL2})}) \cdot V_{\text{in}}(t)$

Eqn. B - 11

Notice the closed loop output voltage at point B in Figure 15 is a replica of input voltage $V_{\text{in}}(t)$ with a factor g_f , and delayed by a factor $(\tau_{A3} + \tau_{DL2})$. Notice also the $V_{\text{IM}}(t)$ is cancelled at the output for a feedforward scheme.

Appendix C

FIGURES

Fig. 1 Frequency Representation of the Nonlinear Voltage Transfer Function $V_{out}(t)$

Fig. 2 Example of $V_{in}(t)$ -- $V_{out}(t)$ relationship

Fig. 3 Generation of IMD products using 2-tone test

Fig. 4 Input-Output Power Relationship

Fig. 5 Feedforward Block Diagram

Fig. 6 OCFF Block Diagram

(b)

Fig. 8 MWO Single Section Stripline Coupler Design:
(a) layout design; (b) 3-dimensional layout.

(a)

(b)

Fig. 8 MWO Single Section Stripline Coupler Design:
(a) layout design; (b) 3-dimensional layout.

(a)

(b)

Fig. 9 MWO stripline coupler result: (a) coupler coupling and insertion loss; (b) coupler phase.

Fig. 10 Coaxial Delay Line Length vs. Delay

Diode_Model
DIODEM1
Is=1.00e-14 Nbv1=1
Rs=4.8 Kf=0
N=1 Af=1
Tt=0 Ffe=1
Cjo=2.3 pF Tnom=27
Vj=3.5 Xti=3
M=1.40 Eg=1.11
Fc=0.5
sr=0
Nr=2
lkf=0
Bv=22
Ibv=1e-05
Nbv=1
Ibv1=0

(a)

(c)

(d)

(e)

Fig. 11 Single Phase Shifter Design: (a) SMV2019 varactor diode model; (b) schematic; (c) test result; (d) control voltage; (e) DLP.

(c)

(d)

Fig. 12 Cascaded Phase Shifter Design: (a) schematic; (b) phase shift vs. control voltage when $V_{ct1} = 1V$; (c) comparison between ideal and cascaded phase shifter; (d) phase range at $V_{ct1} = 0.2V$, $C_{par} = 1pF$ for the other single stage design.

Fig. 13 1st Loop Block Diagram

Fig. 14 1st Loop Suppression with 100 MHz BW Result

Fig. 15 2nd Loop Block Diagram

Fig. 16 A3 TOI and Shunt Gain Relationship

(a)

(b)

Fig. 17 Device Under Test Using HP ADS system amp model: (a) device under test; (b) output spectrum at Pout = 28 dBm.

(a)

(b)

Fig. 18 Simulate coupler used in 5 GHz OCFF system: (a) coupler performance; (b) coupler electrical delay.

Fig. 19 5 GHz OCFF System

Fig. 20 Gshunt vs. A3 TOI

Fig. 21 IM3 when FF is optimized

Fig. 22 Output Spectrum for 2 Different Systems

(a)

freq	TOI	IM3 dBc	TOI rightside	IM3 rightside
<invalid>Hz	50.12	44.54	55.57	55.439

(b)

(c)

Fig. 23 5 GHz OCFF distortion performance for different bandwidth:

(a) $f_1=5.765$ GHz, $f_2=5.785$ GHz; (b) $f_1=5.5$ GHz, $f_2 = 5.7$ GHz;

(c) $f_1=5.2$ GHz, $f_2=5.3$ GHz.

Fig. 24 Overall Efficiency