Adaptive Digital Predistortion Linearization of Frequency Multipliers

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***Abstract*—A novel technique to linearize frequency multipliers for use in high-frequency transmission of digitally modulated signals is presented. Using this technique, a bandpass signal containing both amplitude and phase modulation can be translated without distortion to a higher frequency via nonlinear frequency multiplication. A theoretical analysis is performed to identify the bandpass transformation of the signal envelope in highly nonlinear devices in a manner such that its inverse transformation may be estimated. The theory was validated on a 2.46-GHz Schottky-diode frequency tripler constructed by the authors and on an 820-MHz commercially available frequency doubler. As predicted by the theory, the devices showed highly nonlinear characteristics in terms of their AM/AM, AM/PM, and PM/PM distortions. Adaptive lookup-table- and polynomial-based predistortion systems were designed and constructed to linearize the frequency multipliers. The predistortion results show a fair amount of improvements in adjacent-channel power ratio and error vector magnitude.**

***Index Terms*—Adjacent-channel power ratio (ACPR), frequency multiplier, linearization, nonlinear distortion, predistortion, Schottky-barrier diode.**

I. Introduction

FREQUENCY multiplication is a well-known technique for the generation of high-frequency local-oscillator signals [1]. This technique is preferred when a low phase-noise signal at high frequency is required, but is beyond the practical range of fundamental-mode oscillators and phase-locked loops. Generally, harmonic generation is done by utilizing the nonlinear characteristics of components such as varactor diodes, step-recovery diodes (SRDs), and Schottky-barrier diodes [2], [3]. Active components such as field-effect transistor (FET), MESFET, and high electron-mobility transistor (HEMT) devices are also used to achieve conversion gain [1], [4], [5]. As a result of the highly nonlinear distortion used to generate harmonics, any amplitude modulation (AM) that is present in the signal is also highly distorted. Moreover, phase modulation (PM) undergoes a linear multiplication by the same order as the frequency multiplication. Frequency modulation (FM) may also be used in conjunction with frequency multipliers with a linear scaling in modulation index, but without nonlinear distortion of the baseband envelope. However, the use of AM is often precluded due to the highly nonlinear nature of frequency multipliers. In [6], we demonstrated that frequency multipliers may be used for bandwidth-efficient digitally modulated signals that contain both AM and PM if the distortion is compensated at the device input. In this paper, we expand on that presented in [6] by developing a theory of nonlinear bandpass transformations and show how they may be accurately compensated within their harmonic zones by applying adaptive digital predistortion linearization techniques.

Predistortion linearization is a known technique for reducing distortion in RF power amplifiers (PAs) [7], [8]. The RF PAs used in such systems are generally operated in class AB as a means to trade off linearity for efficiency. Predistortion linearization involves modifying the device input signal to counteract the signal distortion that arises from gain compression (AM/AM distortion) and phase deviation (AM/PM distortion). This is most accurately done by performing mathematical operations on the baseband signals using digital signal processing (DSP). Moreover, the use of digital techniques allows the system to adapt over changes in voltage, temperature, and other environmental factors if a feedback path is incorporated. In the usual architecture, the predistorted signals are converted to analog form and upconverted to the operating frequency, then applied to the PA input. Intermodulation distortion (IMD) products may be suppressed by 10–20 dB using these techniques. Besides PAs, some researchers have also examined the characterization and predistortion of frequency mixers [9]–[12]. As far as frequency multipliers, most of the research has been focused on their optimal design for better conversion loss under saturated operation. Very little attention has been paid to using these as frequency translation devices other than for FM signals. In this paper, we present the first results of applying adaptive digital predistortion to frequency multipliers for the transmission of complex modulated signals. For validation of our theory, two frequency multipliers of different orders were tested. A derivation of bandpass frequency multiplication is developed and presented in Section III, following Section II, which presents some background in practical Schottky-barrier diode frequency multipliers. In addition, a description of the adaptive predistortion algorithms used for the identification and predistortion of frequency multipliers is also given in Section III. Section IV shows the measured results of the predistorted frequency multipliers.

II. Background

Diode circuits are often adopted for harmonic generators at high frequencies. Those diodes can be varactors, SRDs, or Schottky-barrier diodes. Varactor multipliers rely on a nonlinear $C(V)$ characteristic, and are often employed when very low noise multiplication is desired. Multiplication orders

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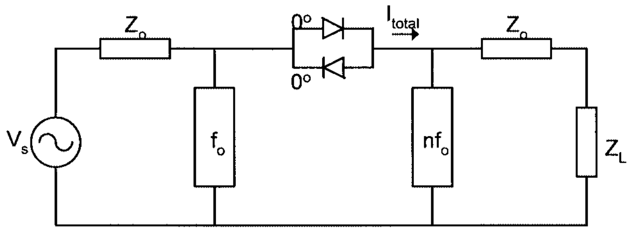


Fig. 1. Circuit diagram of an odd-order frequency multiplier.

of 3–5 may be obtained. However, varactor multipliers are generally narrow-band. SRDs are used for high harmonic generation due to their abrupt reverse recovery properties. They also have the disadvantage of being narrow-band. Schottky multipliers generate harmonics by the exponential behavior of the *I–V* characteristics, and are often restricted to multiplication factors of no more than three. However, Schottky multipliers have significant advantages over reactive multipliers or SRDs in that they are inherently broad-band. Active multipliers generally employ bipolar or field-effect devices, which also have exponential responses over some portion of their usable range. As such, the analysis of Schottky multipliers may be easily extended to active multipliers.

Equation (1) shows the current through a Schottky diode as a function of the applied voltage $V(t)$ [1] as follows:

$$I = I_s\left[\exp(x) - 1\right] \tag{1}$$

where

- x $qV(t)/kT_a$;
- q electron charge;
- k Boltzman constant;
- T_a absolute temperature;
- I_s diode saturation current.

By applying a Fourier series expansion to this equation, we can see the frequency components where a multiplier counts on

$$I = I_s\left[I_o(x) + 2\sum_{1}^{\infty} I_n(x)\cos(n\omega t)\right] \tag{2}$$

where $I_n(x)$ is the modified Bessel function of the first kind, which determines the level of the harmonic components.

Based on the harmonic frequency components, a simple frequency multiplier can be built by connecting a pair of diodes in antiparallel full-wave rectifier fashion, depending on the desired order of harmonic. In addition, input and output circuits must have proper matching circuits that should be seen as open circuits at desired frequencies and as short at all other harmonics. An example of an odd-order diode frequency multiplier is shown in Fig. 1. The circuit can be configured as an even-order multiplier as well by changing the direction of diode currents. In other words, antiphase current of parallel diodes cancels odd harmonics, while the in-phase current of antiparallel diodes cancels even terms. For each case, the current through the load is represented as follows:

$$I_{\text{total}} = 2I_s\left[I_o(x) - 1 + 2\sum_{1}^{\infty} I_{2n}(x)\cos(2n\omega t)\right]: \quad \text{Even order multiplier} \tag{3}$$

$$I_{\text{total}} = 4I_s\left\{\sum_{1}^{\infty} I_{2n+1}(x)\cos\left[(2n+1)\omega t\right]\right\}: \quad \text{Odd order multiplier.} \tag{4}$$

III. Analysis

A. AM/AM, AM/PM, PM/PM Distortions

A complex-modulated bandpass input signal can be represented as

$$x(t) = \text{Re}\left[x_L(t)\cdot e^{j\omega_o t}\right] = A(t)\cos\left(w_o t + \theta(t)\right) \tag{5}$$

where $x_L(t)$ is the complex baseband signal with magnitude $A(t)$ and phase $\theta(t)$. When this signal is fed into a memoryless nonlinear device, its complex output can be expressed by the power series

$$\begin{aligned} y(t) &= \text{Re}\left\{\sum_{n=0}^{\infty} a_n\left[x_L(t)\cdot e^{j\omega_o t}\right]^n\right\} \\ &= \text{Re}\left\{\sum_{n=0}^{\infty} a_n\left[A(t)\cdot e^{j\omega_o t+\theta(t)}\right]^n\right\} \end{aligned} \tag{6}$$

where a_n is a complex power series coefficient.

Note that this is the baseband representation including all the harmonics generated from the device's nonlinearity. In most cases, we are interested only in one bandpass zone: fundamental frequency for PAs and nth zone for frequency multipliers. The relationship between the input and output at the fundamental frequency (first-zone response) has been obtained [13]. For the extraction of the nth zone response, let us take only odd harmonic responses into account. Applying the Chebyshev transform [14] to (6), the nth zone response can be described as follows and distortions can be identified by truncated power series forms of the highest order Q:

$$\begin{aligned} \tilde{y}_n(t) &\cong \text{Re}\left[\sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} \frac{a_{2m+1}}{2^{2m}}\binom{2m+1}{m+\frac{n+1}{2}}|x_L(t)|^{2m-n+1}x_L^n(t)e^{jnw_o t}\right] \\ &= \text{Re}\left[\sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} \frac{a_{2m+1}}{2^{2m}}\binom{2m+1}{m+\frac{n+1}{2}} A(t)^{2m+1}e^{jn\theta(t)}e^{jnw_o t}\right] \\ &= \text{Re}\left[g\left(A(t)\right)e^{j(nw_o t+n\theta(t)+f(A(t)))}\right] \end{aligned} \tag{7}$$

with (8) and (9), shown at the bottom of the following page.

Equations (8) and (9) represent AM/AM and AM/PM distortions, and they are approximated by truncated power series with real coefficients a'_{2m+1} and b'_{2m+1}, respectively. It is noticeable that, in the AM/AM distortion function $g(\cdot)$, the dominant term is generally related to $A(t)^n$, resulting in the power transfer slope of n : 1 in decibel scale.

In addition to the conventional distortions from the device nonlinearity, there exists another distortion arising from the frequency-multiplication process, which we call PM/PM distortion. This can be seen in (7), where $\theta(t)$ is multiplied by n.

In frequency multipliers, noise analysis is often performed by adopting the modulation transfer matrix [15]. We can apply this theory to the case of distortion analysis as well as follows:

$$\begin{bmatrix} m_{\text{out}} \\ \theta_{\text{out}} \end{bmatrix} = \begin{bmatrix} T_{aa} & T_{ap} \\ T_{pa} & T_{pp} \end{bmatrix} \begin{bmatrix} m_{\text{in}} \\ \theta_{\text{in}} \end{bmatrix} + \begin{bmatrix} m_{\text{add}} \\ \theta_{\text{add}} \end{bmatrix} \tag{10}$$

where m and $\theta(t)$ are the AMs and PMs, respectively. The subscripts in and out denote input and output, the subscript add denotes that the noise is an additive component. The basic assumption in this type of analysis is that the noise is stationary and Gaussian. The Ts terms represent AM/AM, PM/AM, AM/PM, and PM/PM modulation transfer coefficients. By expanding this modulation transfer matrix to handle nonlinear distortions as described above, and by limiting the highest order of nonlinear series to Q, (7) and (10) can be combined as

$$\begin{bmatrix} m_{\text{out}}(t) \\ \theta_{\text{out}}(t) \end{bmatrix} = \begin{bmatrix} \mathbf{G} & 0 \\ \mathbf{F} & n \end{bmatrix} \begin{bmatrix} \mathbf{M} \\ \theta_{\text{in}}(t) \end{bmatrix} \tag{11}$$

where

$$\begin{aligned} \mathbf{G} &= \begin{bmatrix} a'_n & a'_{n+2} & \cdots & \cdots & a'_Q \end{bmatrix} \\ \mathbf{F} &= \begin{bmatrix} b'_n & b'_{n+2} & \cdots & \cdots & b'_Q \end{bmatrix} \\ \mathbf{M} &= \begin{bmatrix} A(t)^n & A(t)^{n+2} & \cdots & A(t)^Q \end{bmatrix}^T. \end{aligned} \tag{12}$$

$\mathbf{G}$ and $\mathbf{F}$ are coefficient matrices to represent $g(\cdot)$ and $f(\cdot)$, as in (8) and (9), and $\mathbf{M}$ is the input-magnitude vector. In the distortion analysis of frequency multipliers, we consider the random noise contribution from the device to be negligible compared to the nonlinear distortion components. Thus, the additive noise vector was dropped in (11). This extended transfer matrix enables us to easily express and understand the distortion that arises from nonlinear frequency multipliers. In applying (11), the resulting output nth zone output signal is expressed as

$$\tilde{y}_n(t) = \text{Re}\left[m_{\text{out}}(t)e^{j\theta_{\text{out}}(t)}e^{j\omega_o t}\right]. \tag{13}$$

SectionIII-B discusses the methods used to derive the original input signal from the output described in (13).

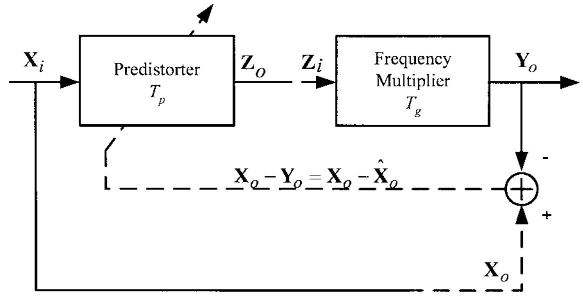


Fig. 2. Block diagram of an adaptive predistorter.

B. Adaptive Predistortion Methods

As shown in Fig. 2 and based on (11), the system's input and output relations are represented by

$$\begin{aligned} \mathbf{Z}_o &= \mathbf{T}_p\mathbf{X}_i: \text{ Predistorter} \\ \mathbf{Y}_o &= \mathbf{T}_g\mathbf{Z}_i: \text{ Nonlinear device} \end{aligned} \tag{14}$$

where

$$\mathbf{T}_p = \begin{bmatrix} \mathbf{G}_p & 0 \\ \mathbf{F}_p & n_p \end{bmatrix} \text{ and } \mathbf{T}_g = \begin{bmatrix} \mathbf{G}_g & 0 \\ \mathbf{F}_g & n_g \end{bmatrix} \tag{15}$$

are the modulation transfer coefficient matrices for the predistorter and nonlinear device, respectively, and

$$\begin{aligned} \mathbf{Z}_o &= \begin{bmatrix} m_Z \\ \theta_Z \end{bmatrix} \\ \mathbf{Y}_o &= \begin{bmatrix} m_Y \\ \theta_Y \end{bmatrix} \\ \mathbf{X}_i &= \begin{bmatrix} \mathbf{M_X} \\ \theta_X \end{bmatrix} \\ \mathbf{Z}_i &= \begin{bmatrix} \mathbf{M_Z} \\ \theta_Z \end{bmatrix} \end{aligned} \tag{16}$$

are output and input vectors, as defined in (11).

The device's modulation transfer matrix $\mathbf{T}_g$ is found by the least squares method from measured data $\mathbf{Y}_o$ and $\mathbf{Z}_i$ that are measured at the input and output of the device

$$\mathbf{T}_g = \mathbf{Z}_i\mathbf{Y}_o^T\left(\mathbf{Y}_o\mathbf{Y}_o^T\right)^{-1}. \tag{17}$$

$$g\big(A(t)\big) = \left|\sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} \frac{a_{2m+1}}{2^{2m}} \begin{pmatrix} 2m+1 \\ m+\dfrac{n+1}{2} \end{pmatrix} A(t)^{2m+1}\right| = \sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} a'_{2m+1}A(t)^{2m+1} \tag{8}$$

$$f\left(A(t)\right) = \arctan\left\{\frac{\text{Im}\left[\sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} \frac{a_{2m+1}}{2^{2m}} \begin{pmatrix} 2m+1 \\ m+\dfrac{n+1}{2} \end{pmatrix} A(t)^{2m+1}\right]}{\text{Re}\left[\sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} \frac{a_{2m+1}}{2^{2m}} \begin{pmatrix} 2m+1 \\ m+\dfrac{n+1}{2} \end{pmatrix} A(t)^{2m+1}\right]}\right\} = \sum_{m=\frac{(n-1)}{2}}^{\frac{(Q-1)}{2}} b'_{2m+1}A(t)^{2m+1} \tag{9}$$

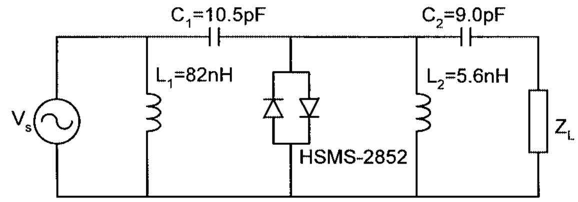


Fig. 3. Schematic of the Schottky diode frequency tripler.

Moreover, the predistortion transfer matrix $\mathbf{T}_p$ is calculated in the same way as follows:

$$\mathbf{T}_p = \mathbf{Z}_o \mathbf{Y}_i^T \left(\mathbf{Y}_i \mathbf{Y}_i^T\right)^{-1} \tag{18}$$

where $\mathbf{T}_p$ represents the inverse of the nonlinear function when $\mathbf{Y}_i$ is the measured output of the nonlinear device formed as the input vector in (16). However, since the statistical distribution of $\mathbf{Y}_i$ is not the same as $\mathbf{X}_i$, which is the input of the predistorter, the $\mathbf{T}_p$ matrix calculated by (18) does not provide the best pre-inverse function for the predistortion system. However, an iterative method may be employed to improve $\mathbf{T}_p$ to find the solution that minimizes the distortion, and provides a better solution for the given measured input and output data.

Another way to find the inverse of a nonlinear device is by iterative generation of a lookup table (LUT) based on measured input and output data. Although it may introduce errors from the quantization of LUT indexes, it can theoretically find an inverse solution for any order of nonlinearity. Moreover, the LUT-quantization error can be reduced by compromising the convergence time and the number of input indexes [8].

IV. Measurement Results

For the validation of our predistortion theory, two different devices were examined. A Schottky diode frequency tripler was designed by the authors for an output at 2.46 GHz, as shown in Fig. 3. The Schottky diodes are Agilent's HSMS-2852. In addition, a Mini-Circuits MK-2 frequency doubler was purchased and tested to produce an output at 820 MHz.

A. Characterization of Devices

A continuous wave input signal at 820 MHz was fed into the frequency tripler and power swept to measure its fundamental- and third-zone output power. The results are shown in Fig. 4. As expected, the third-zone curve shows a 3 : 1 slope over a range of almost 15 dB. Beyond this range, the slope is compressed because of the saturation of current through the Schottky diodes producing relatively strong higher order harmonics. In contrast, nearly a 4 : 1 slope was observed with the Mini-Circuits doubler on the second-zone responses, as shown in Fig. 5, indicating strong nonlinearity was taking place.

For the extraction of AM/AM, AM/PM, and PM/PM responses, a quadrature-modulated signal was fed into the devices, and the output signal was downconverted to an IF frequency. It was then digitized and demodulated. By comparing magnitudes and phases of input and output signals, AM/AM, AM/PM, and PM/PM responses can be found to form (11).

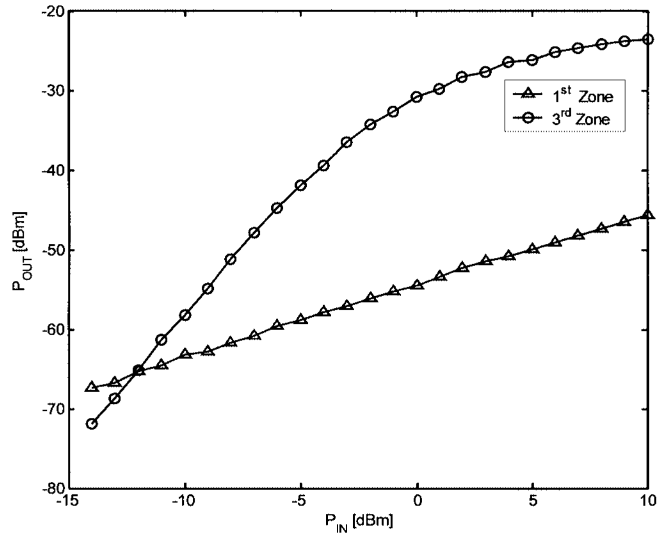


Fig. 4. Measured first- and third-zone transfer characteristics of the diode tripler.

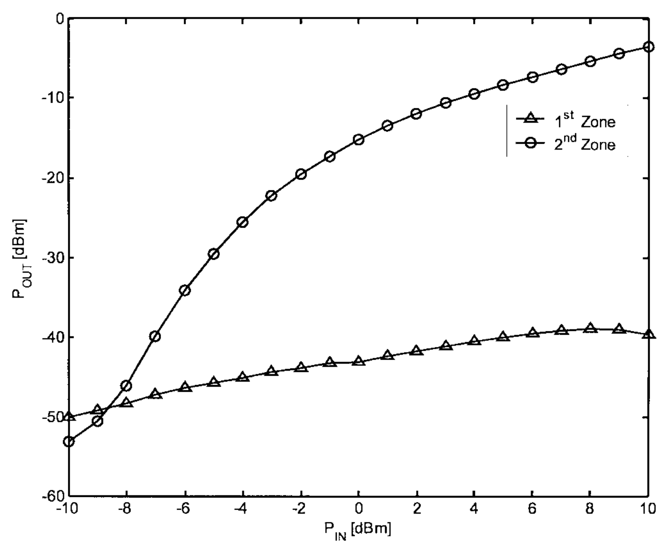
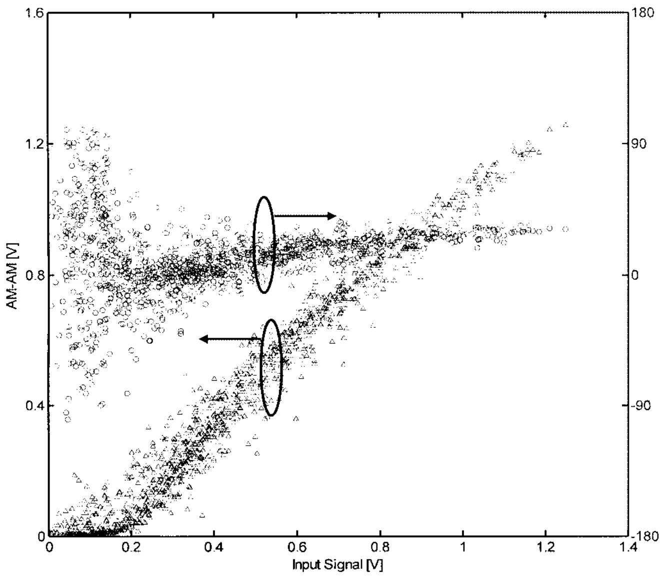
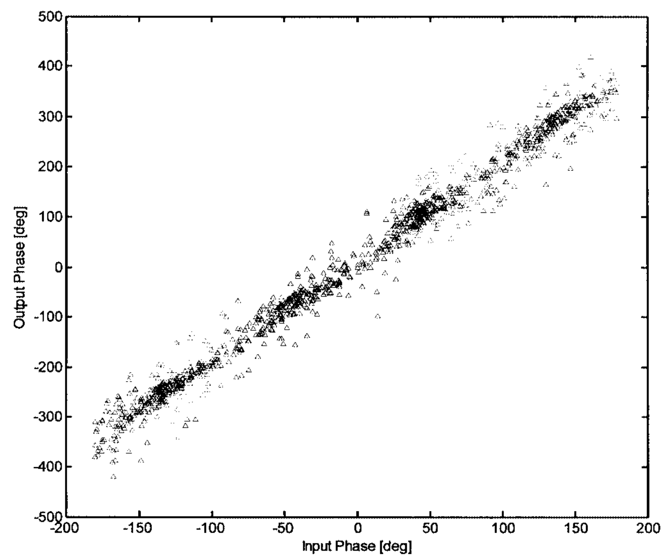


Fig. 5. Measured first- and second-zone transfer characteristics of the diode doubler.

The extracted responses for each device are illustrated in Figs. 6 and 7. From this data, we can clearly see the amplitude and phase nonlinearities are much more severe than those in PAs. Also note in the PM/PM graph that the slope of the asymptotic line represents the phase multiplication. The bunching of data points in 90° increments occurs because the input is an oversampled quadrature phase-shift keying (QPSK) signal. For lower input levels, the output amplitude is nearly zero due to the turn-on voltage of the Schottky diodes. As a result, phase measurements have relatively high variances at lower inputs. While passive diode multipliers serve as useful circuits to validate the theory, the conversion loss is quite high in the range of dominant third- or second-order operation. A more practical approach might be to use active elements (bipolar junction transistors (BJTs) or MESFETs) to achieve some conversion gain in the harmonic generation processes.

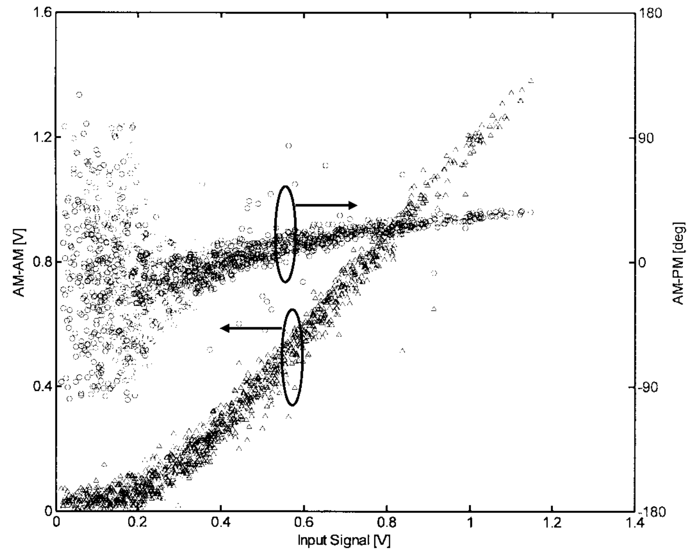


(a)

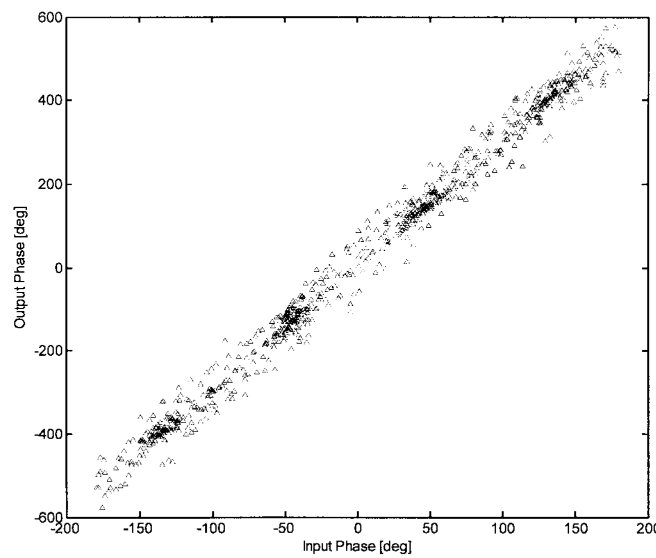


(b)

Fig. 6. Measured characteristics of the frequency doubler. (a) AM/AM and AM/PM responses. (b) PM/PM response.



(a)



(b)

Fig. 7. Measured characteristics of the frequency tripler. (a) AM/AM and AM/PM responses. (b) PM/PM response.

B. Adaptive Digital Predistortion Results

A polynomial-based adaptive digital predistorter was built to find the exact memoryless nonlinear models of devices and to predistort them. As described in the previous section, a complex modulated signal is generated in a PC, and loaded into an Agilent E4432B arbitrary waveform signal generator. For each of frequency multipliers, the predistortion modulation transfer matrix $\mathbf{T}_p$ was extracted from the measured data. Based on $\mathbf{T}_p$, a predistorted signal set was generated, loaded into E4432B, and upconverted to RF. The original input was an IS-95B forward-channel signal at 820 MHz for the tripler and 410 MHz for the doubler. The output signals were taken at 2.46 GHz, and 820 MHz, respectively. To further guarantee the convergence of the polynomial-based predistorter, a LUT-based predistorter was built as well. The LUTs are updated by the least mean-squares method [16]. By comparing the results of two predistorters, we can see that our theory is applicable to both types of predistorters; also be assured that the polynomial can stably express the nonlinearities coming from the frequency-multiplication process. The performance of the predistorter was verified by the output spectrums that are presented in Figs. 8–10. The predistortion result of the tripler is shown in Fig. 8. The original uncompensated output signal was widely spread over 3 MHz due to the phase multiplication, resulting in an adjacent-channel power ratio (ACPR) of almost 4 dBc. The adaptive digital predistortion improved the ACPR to 34 dBc at 885-kHz offset from the center frequency. After predistortion, some residual third-order regrowth still exists due to the truncation in the series. A similar result was observed with the predistortion of the doubler, which is shown in Fig. 9.

The result of LUT-based predistorter of the frequency tripler is in Fig. 10, which shows a 3-dB better suppression of IMDs. The AM/AM LUT entry after predistortion of the frequency

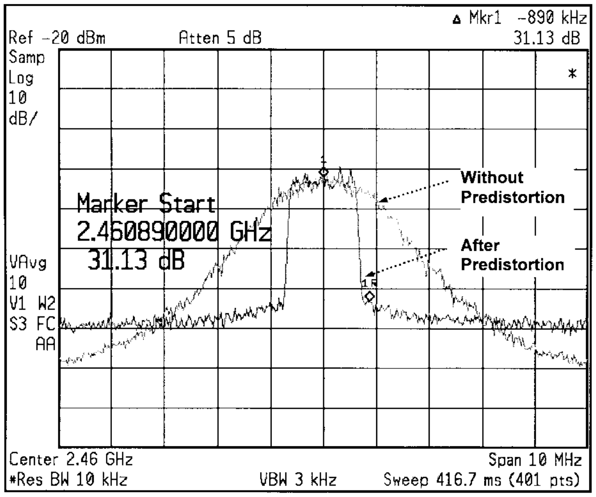


Fig. 8. Measured result of polynomial-based predistortion of a frequency tripler.

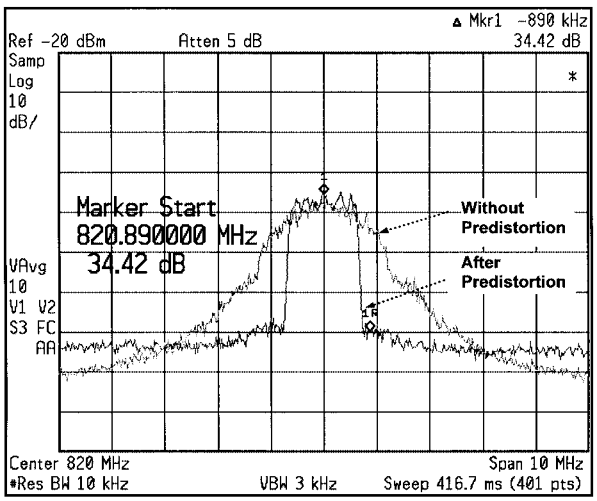


Fig. 9. Measured result of polynomial-based predistortion of a frequency doubler.

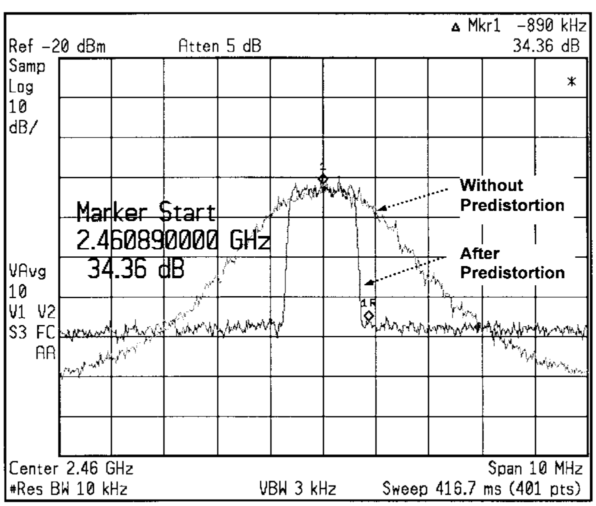


Fig. 10. Measured result of LUT-based predistortion of a frequency tripler.

tripler is expressed by decibel scales in Fig. 11. Notice that the slope is $-2/3:1$ over 10 dB of input scale. This verifies the observation in Fig. 4 that the dominant slope is $3:1$. Therefore, its predistortion gain has to be mainly of the form $\sqrt[3]{1/x^2}$. The

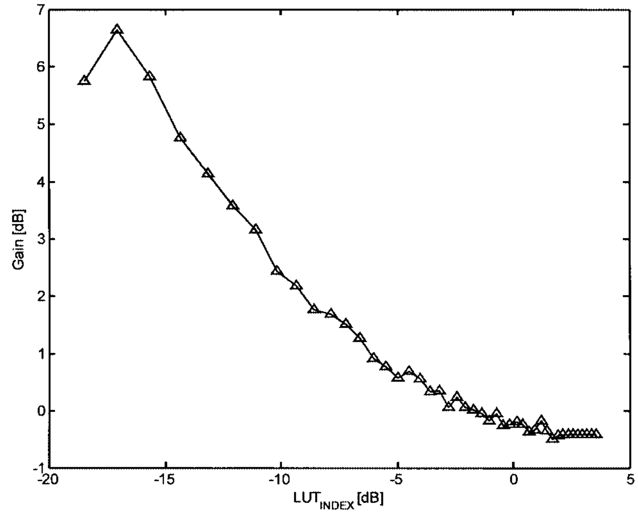


Fig. 11. Measured LUT results for an IS-95B input after adaptive predistortion of a frequency tripler.

in-band error vector magnitude (EVM) was reduced from nearly 100% before predistortion to around 7.3%.

V. Conclusions

This paper has presented a theory and demonstration of predistortion linearization applied to frequency multipliers. The theory was based on baseband predistortion of the input envelope, and the estimated response using the extended transfer matrix was introduced to describe distortions related to frequency multipliers. A simulation model of a frequency tripler implemented on *ADS* showed that a LUT-based digital predistorter can adaptively correct for AM/AM, PM/PM, and AM/PM distortion to improve IMD to better than 40 dBc. To verify this theory, a Schottky diode frequency tripler and doubler at different frequencies were measured for the frequency multiplication of an IS-95B CDMA signal. The frequency tripler showed a $3:1$ input and output power relationship for most of the input range, whereas the doubler showed higher than $2:1$ slope, indicating higher order distortions are affecting the second-zone power transfer characteristics. Moreover, phase multiplications of the baseband signal were observed for both devices. Adaptive digital predistorters based on polynomial series and LUT algorithms were implemented to correct the AM/AM, AM/PM, and PM/PM distortions for the second- and third-zone nonlinearities. The results of the two algorithms, implemented on a test-bed, achieved over 31-dB improvement of the ACPR with an EVM of 7$\sim$8% for all multipliers. Without predistortion, those metrics were approximately 4 dBc and 100%, respectively. The resulting LUT entry from the predistortion of the tripler mainly represents $-2/3:1$ slope in logarithmic scale to compensate for the $3:1$ slope without any predistortion applied to the device. From the results of our research, we can conclude that it is feasible to use predistorted frequency multipliers to transmit digitally modulated signals.

References

[1] E. Camargo, *Design of FET Frequency Multipliers and Harmonic Oscillators*. Boston, MA: Artech House, 1998.

[2] S. A. Maas, "A broadband, planar, monolithic resistive frequency doubler," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1994, pp. 443–446.

[3] J. Zhang and A. V. Räisänen, "Computer-aided design of step recovery diode frequency multipliers," *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 2612–2616, Dec. 1996.

[4] G. Zhao, "The effects of biasing and harmonic loading on MESFET tripler performance," *IEEE Microwave Opt. Technol. Lett.*, vol. 9, pp. 189–194, July 1995.

[5] J. P. Mima and G. R. Branner, "Microwave frequency tripling utilizing active devices," in *42nd Midwest Circuits and Systems Symp.*, vol. 2, 1999, pp. 1048–1051.

[6] Y. Park, K. Low, and J. S. Kenney, "Digital predistortion linearization of frequency multipliers," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2003, pp. 1695–1698.

[7] S. C. Cripps, *Advanced Techniques in RF Power Amplifier Design*. Boston, MA: Artech House, 2002.

[8] J. K. Cavers, "Amplifier linearization using a digital predistorter with fast adaptation and low memory requirements," *IEEE Trans. Veh. Technol.*, vol. 39, pp. 374–382, Nov. 1990.

[9] J. C. Pedro and N. B. Carvalho, "Analysis and measurement of multi-tone intermodulation distortion of microwave frequency converters," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2001, pp. 1671–1674.

[10] T. J. Ellis, "A modified feed-forward technique for mixer linearization," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1998, pp. 1423–1426.

[11] Y. W. Kim, Y. S. Kim, and S. H. Lee, "Linearized mixer using predistortion technique," *IEEE Microwave Wireless Comp. Lett.*, vol. 12, pp. 204–205, June 2002.

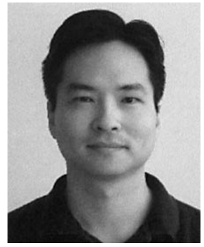
[12] T. Nesimoglu, M. A. Beach, P. A. Warr, and J. R. MacLeod, "Linearised mixer using frequency retranslation," *Electron. Lett.*, vol. 37, pp. 1493–1494, Dec. 2001.

[13] K. G. Gard, H. M. Gutierrez, and M. B. Steer, "Characterization of spectral regrowth in microwave amplifiers based on the nonlinear transformation of a complex Gaussian process," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 1059–1069, July 1999.

[14] N. M. Blackman, "Detectors, bandpass nonlinearities, and their optimization: Inversion of the Chebyshev transform," *IEEE Trans. Inform. Theory*, vol. IT-17, pp. 398–404, July 1971.

[15] A. Riddle, "A CAD program and equations for system phase and amplitude noise analysis," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1989, pp. 359–362.

[16] S. Haykin, *Adaptive Filter Theory*. Upper Saddle River, NJ: Prentice-Hall, 1996.



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