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A MULTICARRIER AMPLIFIER DESIGN LINEARIZED TROUGH SECOND HARMONICS AND SECOND-ORDER IM FEEDBACK

This paper is dedicated to Professor Jovan Surutka on the occasion of his 80th birthday

Bratislav Milovanović, Nataša Maleš-Ilić and Djuradj Budimir

Abstract: A novel linearisation technique for reduction in the first and second kind of the third-order intermodulation products was applied in this paper. The second harmonics and second-order intermodulation products are led from the output to the input of a power amplifier through a feedback loop. The power amplifier including the feedback loop components (bandpass filter, phase shifter and attenuator) was designed as a hybrid microwave integrated circuit by using program ADS. The phase and amplitude of the loop signals are the adjustable parameters. Therefore, a voltage that controls a phase shift of the phase shifter and a control current of a PIN diode in the attenuator circuit were optimised to obtain a reduction in the third-order intermodulation distortion. For three fundamental signals at the power amplifier input, the lowest improvement of 13 dB for the fist and 18 dB for the second kind of the third- order intermodulation product levels was achieved

Key words: Power amplifier, third-order intermodulation, linearization technique, phase shifter, attenuator, bandpass filter.

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B. Milovanović and N. Maleš are with Faculty of Electronic Engineering, Beogradska 14, 18000 Nis, Serbia (e-mail: `[bata,natasam]@elfak.ni.ac.yu`). Dj. Budimir, Department of Electronic Systems, University of Westminster, UK (e-mail: `budimid@cmsa.wmin.ac.uk`).

1. Introduction

In telecommunications systems, the intermodulation (IM) especially the third-order (IM3) generated in-band, has always been of concern, particularly when many channels are simultaneously processed. Many different techniques for IM distortion reduction can be found in literature such as predistortion, feedforward, feedback and combination of them [1]-[2]. However, the application of these techniques requires the circuitry that may be complex, expensive and large in size, as well as limits the usage of active device full capability. In a novel technique for reducing the IM product levels [3], [4], the second harmonics of the input signals are fed together with the fundamental signals to the amplifier input. The injection of the difference frequency between the fundamental signals is another way to reduce IM product levels [5]. Both approaches of a novel technique satisfy the reduction of IM product levels without affecting the fundamental signal power levels. Additionally, the required circuitry is simple, inexpensive and small in size. The authors of this paper have investigated in [6,7] the effects of the injection of carrier second harmonics on the intermodulation in a microwave power amplifier with two-tone test at amplifier input. The influence of the second harmonics as well as the second-order intermodulation products (IM2) at frequencies which are the sum of pairs of the fundamental signal frequencies (frequency summation technique) to the first and second kind of third-order IM was reported in [8]. The published results were based on these signals' generation and their injection into the amplifier input together with three fundamental signals.

In our work, all these signals are extracted from a non- linear power amplifier output, in contrast to above mentioned approach, and returned to the amplifier input through the feedback loop. The loop components (band-pass filter, phase shifter and attenuator) were designed for the application in a hybrid MIC of power amplifier. Simulation and design of a single stage power amplifier as well as the feedback loop components were performed by the microwave circuit simulator Advanced Design System (ADS).

2. Analysis

The proposed technique uses the amplifier non-linear characteristic to generate the second third-order IM signals (the first and second kind) that are used to cancel the original third-order IM products at the output. An expression for the non-linearity of the active device (MESFET) is represented by a three term Taylor's series connecting the input voltage, V_{in} with the output current, I_{out} and the transconductance, g_m regarded as the dominant non-linearity

$$I_{out} = g_{m_1} V_{in} + \frac{1}{2} g_{m_2} V_{in}^2 + \frac{1}{6} g_{m_3} V_{in}^3 \tag{1}$$

with $g_{m_1} = \mathrm{d}I_d/\mathrm{d}V_g$, $g_{m_2} = \mathrm{d}^2 I_d/\mathrm{d}V_g^2$ and $g_{m_3} = \mathrm{d}^3 I_d/\mathrm{d}V_g^3$.

The second harmonics as well as the second-order intermodulation signals are the products of amplifier non- linearity. Therefore, these products generated at the output are led trough feedback loop to the input of power amplifier.

A three-tone injection of the fundamental signals at the frequencies ω_1, ω_2 and ω_3 with amplitudes V_{ω_1}, V_{ω_2} and V_{ω_3} respectively, together with their second harmonics at the frequencies $2\omega_1$, $2\omega_2$ and $2\omega_3$ with amplitudes $V_{2\omega_1}$, $V_{2\omega_2}$ and $V_{2\omega_3}$ and phases $\varphi_{2\omega_1}$, $\varphi_{2\omega_2}$ and $\varphi_{2\omega_3}$ can be expressed as

$$\begin{aligned} V_{in} =& V_{\omega_1} \cos(\omega_1 t) + V_{\omega_2} \cos(\omega_2 t) + V_{\omega_3} \cos(\omega_3 t) \\ &+ V_{2\omega_1} \cos(2\omega_1 t + \varphi_{2\omega_1}) + V_{2\omega_2} \cos(2\omega_2 t + \varphi_{2\omega_2}) \\ &+ V_{2\omega_3} \cos(2\omega_3 t + \varphi_{2\omega_3}) \end{aligned} \tag{2}$$

The source signals and their second harmonics interact as the result of amplifier non-linearity. All relevant frequency components at the output of the amplifier can be obtained by substituting equation (2) into equation (1). The expression valid for the first kind of IM3 product at the frequency $(2\omega_2 - \omega_1)$ is

$$\begin{aligned} I_{out_{(2\omega_2-\omega_1)}} =& \frac{3}{4} V_{\omega_1} V_{\omega_2}^2 g_{m_3} \cos(2\omega_2 t - \omega_1 t) \\ &+ V_{\omega_1} V_{2\omega_2} g_{m_2} \cos(2\omega_2 t - \omega_1 t + \varphi_{2\omega_2}) \\ &+ \frac{3}{2} V_{\omega_1} V_{2\omega_1} V_{2\omega_2} g_{m_3} \cos(2\omega_2 t - \omega_1 t + \varphi_{2\omega_2} - \varphi_{2\omega_1}) \end{aligned} \tag{3}$$

The first term in (3) relates to the third-order IM product caused by the interaction between fundamental signals. The interaction between the fundamental signals and second harmonics results in the additional signals at the output of the amplifier at the third-order IM frequencies, (the second and third terms in (3)). Therefore, by a proper selection of phase and amplitude of the second harmonics, it is possible to make the third-order IM products produced by the second harmonics out of the phase and equal in the amplitude with the original third-order product. Similar conditions are valid for the cancellation of the other IM3 components at frequencies ($2\omega_1 - \omega_2$, $2\omega_1 - \omega_3$, $2\omega_2 - \omega_3$, $2\omega_3 - \omega_1$ and $2\omega_3 - \omega_2$).

Theoretical investigations [8] imply that the single second harmonic injection does not provide the reduction of the second kind of IM3.

The expression for the injection of the second-order IM signals at frequencies which are the sum of the fundamental signal frequencies is:

$$\begin{aligned} V_{in} =& V_{\omega_1}\cos(\omega_1 t) + V_{\omega_2}\cos(\omega_2 t) + V_{\omega_3}\cos(\omega_3 t) \\ &+ V_{\omega_{12}}\cos(\omega_1 t + \omega_2 t + \varphi_{\omega_{12}}) + V_{\omega_{23}}\cos(\omega_2 t + \omega_3 t + \varphi_{2\omega_{23}}) \\ &+ V_{\omega_{13}}\cos(\omega_1 t + \omega_3 t + \varphi_{\omega_{13}}) \end{aligned} \tag{4}$$

where the injected signals $(\omega_1+\omega_2)$, $(\omega_2+\omega_3)$ and $(\omega_1+\omega_3)$ have amplitudes $V_{\omega_{12}}$, $V_{\omega_{23}}$ and $V_{\omega_{13}}$ and phases $\varphi_{\omega_{12}}$, $\varphi_{\omega_{23}}$ and $\varphi_{\omega_{13}}$, respectively. The additional signals at the second kind IM3 frequencies at output of the amplifier are the results of interaction between the source signals (ω_1, ω_2 and ω_3) and the injected IM2 signals. The expressions for the first kind of IM3 product at $2\omega_1 - \omega_2$, $2\omega_1 - \omega_3$, $2\omega_2 - \omega_1$, $2\omega_2 - \omega_3$, $2\omega_3 - \omega_1$ and $2\omega_3 - \omega_2$, contain the additional term caused by the injection of the second-order IM products, which exhibits a minimum, what is the reason of the small reduction in the IM3 first kind. The second kind of IM3 at the frequency $(\omega_1 + \omega_2 - \omega_3)$ is expressed as given in (5) where the first term relates to the interaction between fundamental signals only. The original signals and the injected second-order IM signals interact and produce additional third-order terms of the second kind. Therefore, a complete elimination of the second kind of IM3 products can be achieved by a proper selection of phases and amplitudes of the injected signals. Similar conditions are found for the other products $(\omega_1 - \omega_2 + \omega_3)$ and $(\omega_3 + \omega_2 - \omega_1)$. The analysis shows that the injection of the second harmonic reduces the first kind of IM3 and the injection of the second- order IM products reduces the second kind of IM3.

$$\begin{aligned} I_{out_{(\omega_1+\omega_2-\omega_3)}} =& \frac{3}{2}V_{\omega_1}V_{\omega_2}V_{\omega_3}g_{m_3}\cos(\omega_1 t + \omega_2 t - \omega_3 t) \\ &+ \frac{3}{2}V_{\omega_3}V_{\omega_{21}}g_{m_3}\cos(\omega_1 t + \omega_2 t - \omega_3 t + \varphi_{\omega_{21}}) \\ &+ \frac{3}{2}V_{\omega_1}V_{\omega_{21}}V_{\omega_{31}}g_{m_3}\cos(\omega_1 t + \omega_2 t - \omega_3 t + \varphi_{\omega_{21}} + \varphi_{\omega_{31}}) \\ &+ \frac{3}{2}V_{\omega_2}V_{\omega_{21}}V_{\omega_{32}}g_{m_3}\cos(\omega_1 t + \omega_2 t - \omega_3 t + \varphi_{\omega_{21}} + \varphi_{\omega_{32}}) \end{aligned} \tag{5}$$

3. Feedback Loop Components

The power amplifier circuit with feedback loop is presented in Fig. 1. A design was applied on the substrate characterised by following parameters ε_r=4.3, H=0.635 mm, t=0.004 mm. Whole power amplifier circuit was analyzed with the included effects of the microstrip discontinuities and transistor biasing circuits.

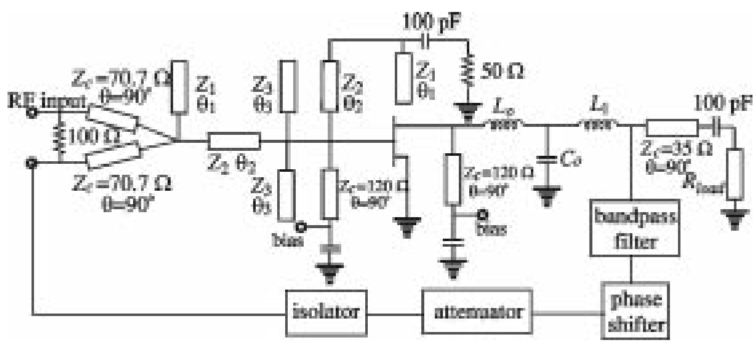


Fig. 1. Power amplifier with the second harmonics and second-order IM feedback loop.

The first component in the feedback loop is a bandpass filter in order to extract only the second harmonics and IM2 products from the output of the power amplifier. The capacitive-gap coupled bandpass filter [9] that can be conveniently fabricated in microstrip, is designed at 5.3 GHz centre frequency with 20 % bandwidth and 0.5 dB equal-ripple response, with 3 sections. The application of this type of bandpass filters as the feedback loop component of the power amplifier does not affect the complete amplifier characteristics at the frequencies out of filter bandpass. However, the filter was designed with a wider bandwidth than the extraction of the second harmonics and the second-order IM really requires. Namely, wider bandpass range and smaller number of filter sections provide a smaller variation of the bandpass filter phase characteristic in terms of frequency. Such characteristic is needed so that less differences between the phases of the second harmonics and second-order IM may be achieved.

A 360° reflection-type analog phase shifter with a single 90° branch-line coupler [10] was designed. The phase shift in this phase shifter shown in Fig. 2, is produced by reflecting the incident wave from a varactor diode whose capacitance varies according to the bias voltage. As insertion loss strongly depends on impedance Z_1 (Fig. 2), its appropriate value to the reflection load (two varactors in series with a shorted stub and separated by quarterwave transmission line) is provided by the impedance transformer. GaAs hyperabrupt tuning varactor type MA-46553 was used for simulation. Varying the controlled voltage between 2 and 20 V, a full phase shift of 360° is achieved, accompanied with the phase shifter attenuation of 3.4 dB, at the frequencies of the fundamental signal second harmonics and IM2 products.

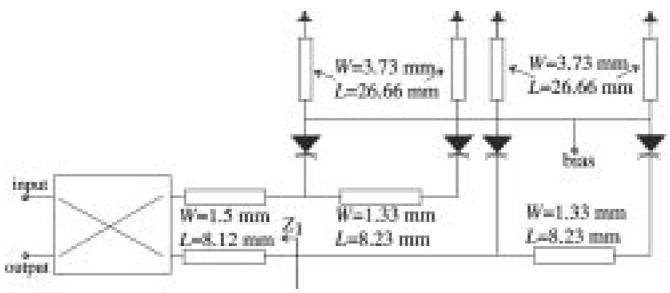


Fig. 2. Phase shifter.

The realtiv phase shift characteristics of the phase shifter over the frequency range 5-5.2 GHz for the control voltage values Vvar are shown in Fig. 3.

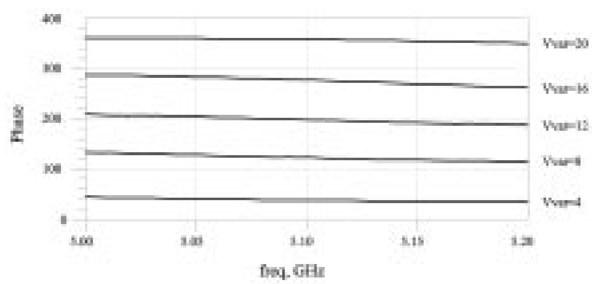


Fig. 3 Relativ phase shift of reflection type phase shifter.

Regarding the results shown in Fig. 3, it can be seen almost constant relative phase shift characteristics over the frequency range 5-5.2 GHz.

PIN diode attenuator structure shown in Fig. 4 accomplishes an appropriate attenuation of the second harmonics and the second-order IM. For simulation purposes, HP hermetic PIN diode for microstrip attenuators was used. Changing a diode forward current from 0.01 to 100 mA, the resistance of the intrinsic region of the diode is varied, providing the control of attenuation with the bias point. The input and output of the attenuator are at the input and isolated ports of the 3 dB Lange-coupler selected in order to attain better performances over wider frequency range.

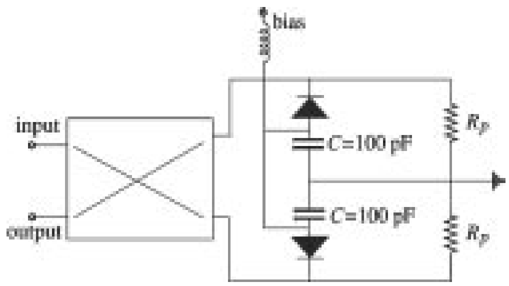


Fig. 4. Variable attenuator.

Simulated results show that the attenuation controlled by the bias current varies from 0.5-33 dB. The attenuation characteristics of variable attenuator in the frequency range 5-5.4 GHz, that relate to PIN diode intrinsic resistance Rin are represented in Fig. 5. This figure shows a flat attenuation characteristic over the frequency range 5-5.4 GHz for all values of attenuation.

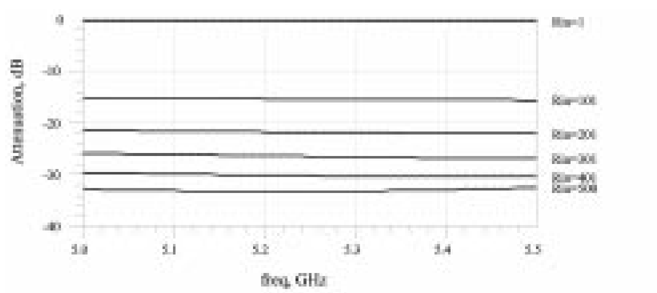


Fig. 5 Attenuation of variable atteniator.

The lack of appropriate element in ADS library limits the design of the isolator in the feedback loop, and an ideal library element was used for this component.

4. Numerical Results

Three main input fundamental signals were chosen at frequencies 2.5 GHz, 2.51 GHz and 2.522 GHz, and their input power levels are –2 dBm. Non-linear Curtice's cubic model was used in the CAD simulation for MESFET modeling.

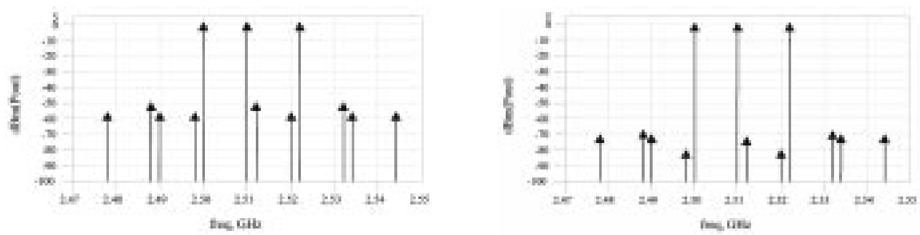
The fundamental input signals and returned signals of the second harmonics and the second-order IM are combined at the amplifier input over Wilkinson's combiner, Fig. 1.

An input broadband matching circuit was designed combining two bisected π matching sections [6] which, transformed in the circuit suitable for realization in a microstrip technique, are shown in Fig. 1. Input reflection coefficient less than –10 dB in the frequency range 2-9 GHz as well as unconditionally stable operation of MESFET were achieved by the proposed input matching circuit.

The concept of output matching circuit is shown in Fig. 1. The values of L_0, C_0, L_1, and transformer impedance were changed by using optimisation facility of program ADS in order to accomplish desired performance of fundamental signals' output power of approximately 0.7 dBm and the lowest power of the third-order IM signals, for the first kind –56 dBm and for the second kind –50 dBm. Also, it is desirable to attain as much as possible less discrepancies between IM3 power levels of a certain kind.

The spectrum for the bias point $V_{gs} = -0.4$ V and $V_{ds} = 3$ V, obtained at the amplifier output without applying our technique is shown in Fig. 6(a). It includes fundamental signals, the first kind of the third-order IM products at 2.478, 2.49, 2.498, 2.52, 2.534 and 2.544 GHz and the second kind of IM3 products at 2.488, 2.512 and 2.532 GHz. When the second harmonics and the second-order IM products were returned from the output to the input of the power amplifier, CAD optimisation was used to find the right phase and amplitude of these signals in order to reduce the IM3 products by keeping the fundamental signal power levels constant. The spectrum obtained after simulation are shown in Fig. 6(b).

The results obtained during analyses show that a maximum reduction of approximately 40 dB of each IM3 product can be obtained separately without reduction of the fundamental signals. The values of the phase shift and attenuation at which maximum reduction is obtained are different for each IM3 product. The analysis shows that all components of a particular kind of feedback signals must be controlled within a fraction of dB in ampli-



(a) (b)

Fig. 6. The simulated fundamental powers and third-order IM powers (a) Before employing; (b) After employing the technique.

tude and of a few degrees in phase in order to attain the maximal reduction in all IM3 products. On the other hand, as these products have slightly different amplitudes and phases, observing independently for two IM3 kinds, it is difficult to obtain a maximal reduction in all IM3 products with the same value of the amplitude and phase adjustment. The results shown in Fig. 6 refer to the compromise between the maximum reduction in each IM3 signal obtained separately, yielding to the lowest improvement of 13 dB for the fist kind and 18 dB for the second kind of IM3. These results were achieved with phase shift of 44° and attenuation of 0.5 dB observed at 5 GHz.

Monte Carlo analysis shows that, altering the attenuation for ±2 dB, the reduction in IM3 decreases for 5 dB, while the change of the phase shift by ±8° will lead to the 5 dB lower reduction.

5. Conclusion

In order to reduce the third-order intermodulation products of power amplifier, a novel linearisation technique was applied. The second harmonics and second-order intermodulation products of the fundamental signals were injected through the feedback loop of power amplifier. The proposed technique uses the amplifier non-linear characteristic to generate the first and second kind of the third-order IM signals that are used to cancel the original third-order IM products at the output. The earlier published results referring to the same novel technique approach are based on the power amplifier simulation with the additional second harmonics and second-order IM signals generated and injected together with fundamental signals. In this paper, non- linear signals are led from the amplifier output to its input through the feedback loop. The loop components (bandpass filter, phase shifter, attenuator) as well as a single stage amplifier circuit were designed as a hybrid microwave integrated circuit in a microstrip technique by using program ADS. Adjusting the phase and amplitude of the loop signals the lowest reduction obtained in the first and the second kind of the third-order intermodulation products are 13 dB and 18 dB, respectively for fundamental signals at 2.5, 2.51 and 2.522 GHz. Simulated results referring to the designed phase shifter and variable attenuator show almost constant characteristics of the relative phase shift over frequency range 5-5.2 GHz and straight attenuation characteristics for 5-5.4 GHz frequency range.

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