DESIGN OF HYBRID MIC POWER AMPLIFIER LINEARIZED TROUGH SECOND HARMONICS FEEDBACK

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Abstract - **A novel linearisation technique for reduction in the third-order intermodulation distortion products, with the injection of the second harmonics through a feedback loop of a power amplifier was applied in this paper. The power amplifier including the feedback loop components (band pass filter, phase shifter, attenuator) was designed as a hybrid microwave integrated circuit by using program Libra. The adjustable parameters are the phase and amplitude of the loop signals. 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. The achievable improvement was found to be 28 dB for the case of two fundamental signals at the power amplifier input.**

I. INTRODUCTION

In multicarrier communications systems, the intermodulation distortion (IMD) products, especially the third-order (IMD3), represent the most serious problem. Many different techniques for IMD 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 IMD 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 IMD product levels [5]. Both approaches of a novel technique satisfy the reduction of IMD product levels without affecting the fundamental signal power levels. Additionally, the required circuitry is simple, inexpensive and small in size. But, a novel technique based on difference frequency injection is unsuitable for practical applications because a feedback loop must have a sufficient bandwidth to yield the distortion improvement over a range of carriers' spacing used.

This work extends previous analyses of a multicarrier amplification. The effects of the injection of carrier second harmonics on the intermodulation in a microwave power amplifier were investigated. The published results preceding this paper were based on two ways for the injection of the second harmonic signals. In one approach the second harmonics were led to the amplifier input across the feedback loop which components (band pass filter, phase shifter, attenuator and isolator) were modeled by ideal elements from the library of commercial programs such as Libra or MDS. In the other, instead of feeding back the second harmonics, they were generated and injected into the amplifier input together with fundamental signals in the simulations as well as in the experiments. In our work, second harmonics are extracted from a nonlinear power amplifier output, and returned to the amplifier input through the feedback loop whose components (band pass filter, phase shifter and attenuator), in contrast to above mentioned approaches, 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 Libra. The lack of appropriate element in Libra library limits the isolator design, and an ideal library element was used for this component.

II. ANALYSIS

The proposed technique uses the amplifier nonlinear characteristic to generate a second third-order IMD signal that is used to cancel the original third-order IMD product at the output.

An expression for the non-linearity of the active device (MESFET) is represented by a three term Taylors series connecting the input voltage, V_{in} with the output current, I_{out} and the transconductance, g_m regarded as the dominant non-linearity:

$$I_{out} = g_{m_1} V_{in} + \frac{1}{2} g_{g_2} V_{in}^2 + \frac{1}{6} g_{m_3} V_{in}^3 \qquad (1)$$

with $g_{m_1} = \dfrac{dI_d}{dV_g}$, $g_{m_2} = \dfrac{d^2 I_d}{dV_g^2}$ and $g_{m_3} = \dfrac{d^3 I_d}{dV_g^3}$.

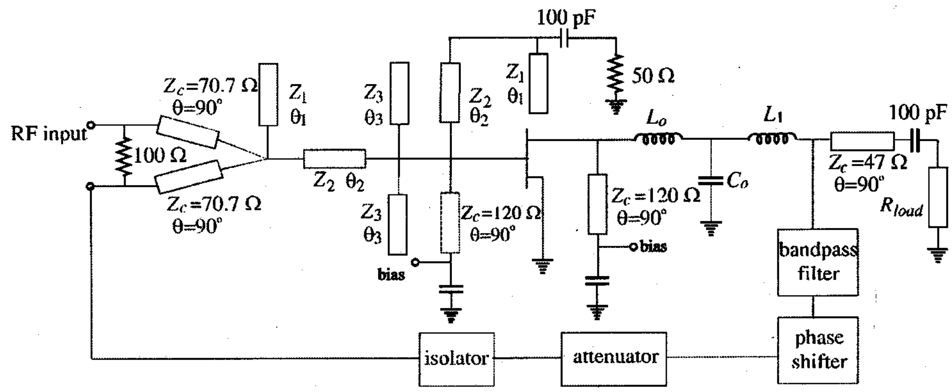

Fig. 1. - Power amplifier with the second harmonics feedback loop

Second harmonics are the products of amplifier non- linearity. So, the second harmonic products generated at the output are led trough feedback loop to the input of power amplifier.

A two-tone injection of the fundamental signals at the frequencies ω_1 and ω_2 with amplitudes $A_{\omega 1}$ and $A_{\omega 2}$ respectively, together with their second harmonics at the frequencies $2\omega_1$ and $2\omega_2$ with amplitudes $A_{2\omega 1}$ and $A_{2\omega 2}$ and phases $\phi_{2\omega 1}$ and $\phi_{2\omega 2}$ can be expressed as:

$$V_{in}=A_{\omega_1}\cos(\omega_1 t)+A_{\omega_2}\cos(\omega_2 t)$$

$$A_{2\omega_1}\cos(2\omega_1 t+\phi_{2\omega_1})+A_{2\omega_2}\cos(2\omega_2 t+\phi_{2\omega_2}) \quad (2)$$

Interaction between the source signals and their second harmonics is the results of amplifier non-linearity, too. Substituting equation (2) into equation (1), all the relevant frequency components at the output of the amplifier can be obtained. For the IMD3 product at the frequency $(2\omega_2-\omega_1)$, the following expression is valid:

$$I_{out_{(2\omega_2-\omega_1)}}=\frac{3}{4}A^2_{\omega_2}A_{\omega_1}g_{m_3}\cos(2\omega_2 t-\omega_1 t)$$

$$+A_{\omega_1}A_{2\omega_1}g_{m2}\cos(2\omega_2 t-\omega_1 t+\phi_{2\omega_2})$$

$$+\frac{3}{2}A_{\omega_1}A_{2\omega_1}A_{\omega_2}g_{m3}\cos(2\omega_2 t-\omega_1 t+\phi_{2\omega_2}-\phi_{2\omega_1}) \quad (3)$$

Only interaction between fundamental signals causes the components of the third-order IMD products, (the first term in (3)). The result of interaction between the fundamental signals and second harmonics are the additional signals at the output of the amplifier at the third-order IMD frequencies, (the second and third terms in (3)).

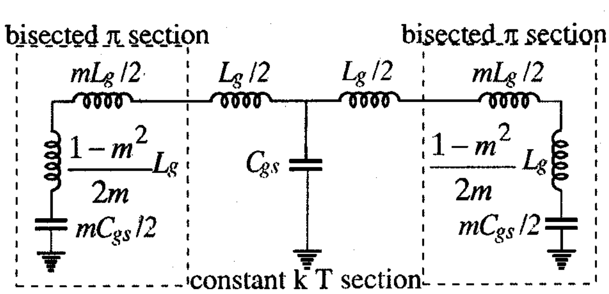

Fig. 2. - Input matching circuit

Therefore, by a proper selection of phase and amplitude of the second harmonics, it is possible to make the third-order IMD products produced by the second harmonics out of the phase and equal in the amplitude with the original third-order product. Similar conditions are found for the cancellation of the other IMD3 component at frequency $(2\omega_1-\omega_2)$.

III. POWER AMPLIFIER WITH 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. In power amplifier simulation NEC MESFET from Libra library denoted as NE710 was used.

An input matching circuit was designed as a bisected π matching section regarding that input equivalent transistor circuit is treated as a shunt capacitance in series with a low resistance.

In order to obtain circuit suitable for realization in microstrip technique as shown in Fig. 1, Richards transformation and Kurods identities were used. Input matching was performed to obtain the satisfactory input reflection coefficient and an unconditionally stable operation of MESFET.

The concept of output matching circuit shown in Fig. 1, was applied to give the highest output power of fundamental signals and the lowest power of IMD3 signals as well as to attain as much as possible less discrepancies between IMD3 power levels.

The influence of the transistor biasing circuits was included into analyses. Whole power amplifier circuit was analyzed with included effects of microstrip discontinuities.

With the aim to select only the second harmonic signals at the output of the power amplifier, the first component in the feedback loop must be a bandpass filter. The bandpass filter that can be conveniently fabricated in microstrip is the capacitive-gap coupled resonator filter [6] designed at 5.3 GHz centre frequency with 20% bandwidth and 0.5 dB equal-ripple response, with 3 sections. This type of bandpass filters is suitable for applica-

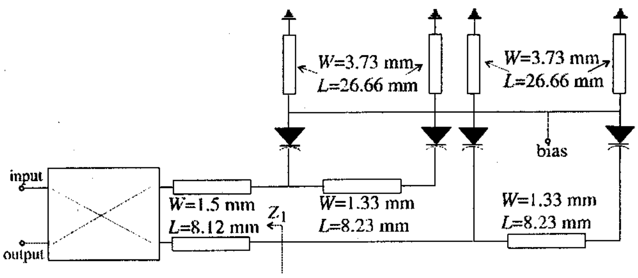

Fig. 3. - Phase shifter

tion as the feedback loop component of the power amplifier, because its input impedance characteristics is such that there is no influence to the complete amplifier characteristics at the frequencies out of filter bandpass.

The filter is designed with much wider bandwidth than the extraction of the second harmonics really requires because of the bandpass filter phase characteristic. Namely, the slop of this parameter, in terms of frequency, decreases considerably with wider bandpass range and smaller number of filter sections. Such characteristic is needed so that smaller differences between the phases of the second harmonics may be achieved.

Neglecting this resistance, the equivalent capacitance was extended by $L_g/2$ ($L_g = 50^2 C_{gs}$) as shown in Fig. 2, forming constant k T section known in a filter design. Therefore, bisected π matching section with m=0.6 can be connected at both sides of *k T* section for broadband matching, Fig 2.

A 360^0 reflection-type analog phase shifter with a single 90^0 branch-line coupler [7] was designed. The phase shifter circuit is shown in Fig. 3. The phase shift in this phase shifter is produced by reflecting the incident wave from a varactor diode whose capacitance varies according to the bias voltage. The symmetric impedance variation over the whole voltage control range is produced by a shorted microstrip line connected in series with the varactor. A reflection load proposed in [7], consisting of two varactors in series with a shorted stub and separated by quarterwave microstrip line is used to achieve the full 360^0 phase shift with an acceptable insertion loss. As insertion loss strongly depends on impedance Z_1 (Fig. 3), its appropriate value to the reflection load is provided by the impedance transformer. The direct and the coupled ports of 90^0 branch-line coupler are connected to

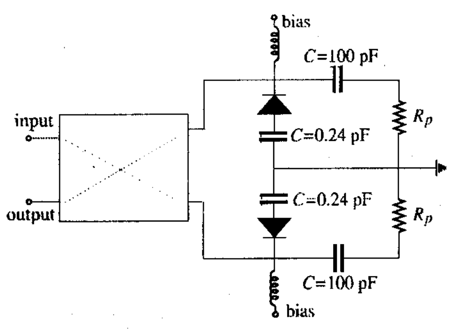

Fig. 4. - Variable attenuator

two sets of impedance transformer and reflection load. 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 360^0 is achieved at the frequencies of the fundamental signal second harmonics.

PIN diode attenuator shown in Fig. 4 accomplishes an appropriate attenuation of the second harmonics. For simulation purposes, HP packaged PIN diode 5082-3001 for hybrid MIC 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. A capacitance in series with the forward biased diode was chosen that, together with the reactance of the diode, represents a resonant circuit at the design frequency of the attenuator. The input and output of the attenuator are input and isolated ports of the coupler selected to be 3 dB Lange- coupler in order to attain better performance over wider frequency range. Simulated results show that the attenuation controlled by the bias current varies from 0.8-18.5dB.

Returned signals of the second harmonics are combined with fundamental input signals over Vilkinsons combiner, Fig. 1.

IV. NUMERICAL RESULTS

The chosen frequencies of two main input fundamental signals are 2.5 GHz and 2.51 GHz and their input power levels are 2 dBm. In the CAD simulation, non-linear Curtices cubic model was used for MESFET modeling. Input reflection coefficient less than 10 dB was achieved by proposed input matching circuit (Fig. 1) in frequency range 2-9 GHz. The values of L_0, L_1, C_0 and transformer impedance, Fig. 1, were changed by using optimisation facility of program Libra in order to accomplish desired performance of fundamental signals output power of approximately 3 dBm and the lowest power of third order IMD signals of 57dBm.

The realtiv phase shift characteristics of the phase shifter over the frequency range 5-5.2 GHz for the control voltage values of 4, 9, 14 and 20 V, denoted as PHR- control voltage, are shown in Fig. 5. The attenuation characteristics of variable attenuator in the same frequency range, that relate to PIN diode intrinsic resistance of 1, 25, 100 and 10000 Ω are represented in Fig. 6, with an appropriate denotation ATEN-resistance value. Regarding the results shown in Fig. 5, it can be seen almost constant relative phase shift characteristics over the frequency range 5-5.2 GHz. Fig. 6 shows a slight deviation of about 1 dB in attenuation over the same frequency range for 0.8-6 dB attenuation and almost constant characteristics for higher values of attenuation.

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. 7a). It includes fundamental signals and the third-order IMD products at 2.49 GHz and 2.52 GHz. When

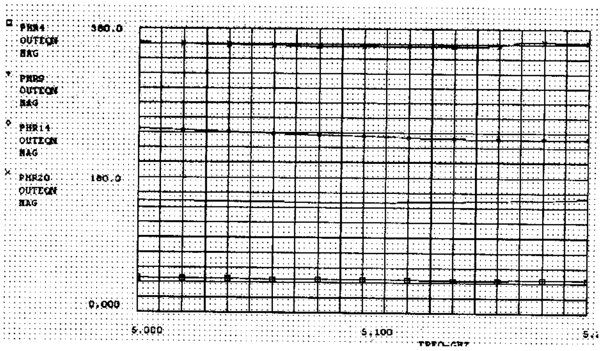

Fig. 5 - Relativ phase shift of reflection type phase shifter

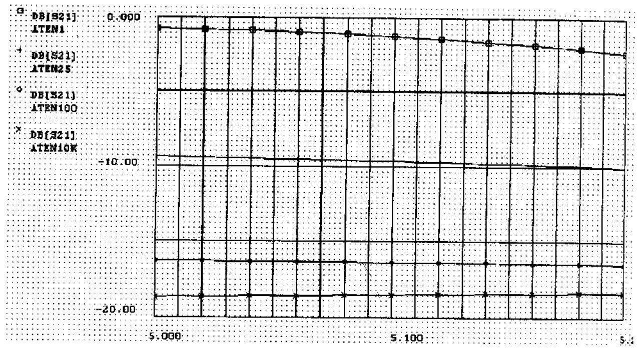

Fig. 6 - Attenuation of variable atteniator

the second harmonics 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 IMD3 products by keeping the fundamental signal power levels constant. The spectrum obtained after simulation are shown in Fig. 7b).

The results obtained during analyses show that a maximum reduction of approximately 40 dB of each IMD3 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, do not coincide. The second harmonic spectral components must have the approximately the same power level, so that all of them can be controlled within a fraction

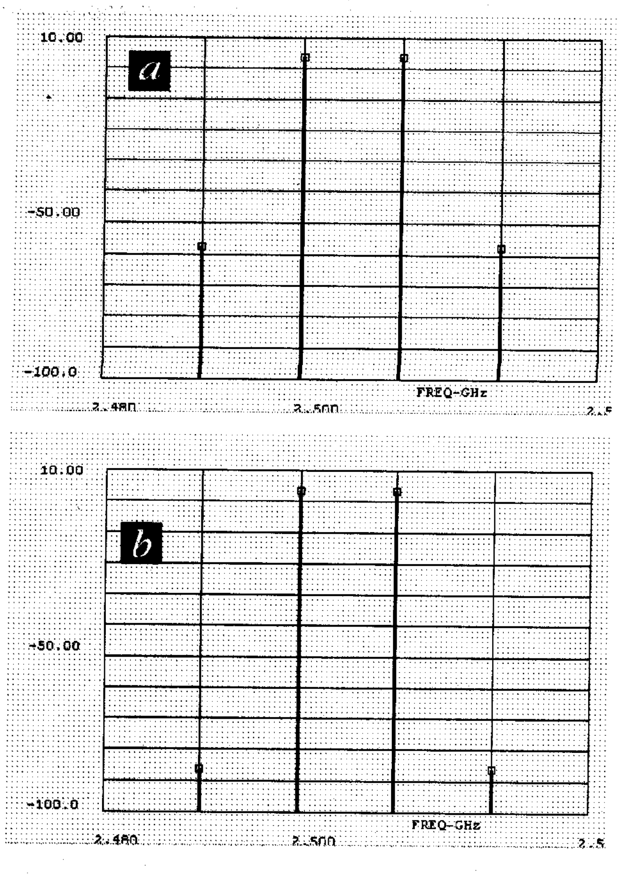

Fig. 7. - The simulated fundamental powers and third-order IMD powers
a) Before employing the technique;
b) After employing the technique.

of dB in amplitude and of a few degrees in phase in order to attain maximum reduction in IMD3 products. On the other hand, it is difficult to obtain a maximum reduction in all IMD3 products with the same value of amplitude and phase adjustment due to; these products have slightly different amplitudes and phases. The results shown in Fig. 7 refer to the compromise between the maximum reductions in each IMD3 signal obtained separately, yielding 28 dB improvements in both. These results were achieved with phase shift of 215.4^0 and attenuation of 0.9 dB.

V. CONCLUSION

In order to reduce the third-order intermodulation products of power amplifier, a novel linearisation technique was applied. The second harmonics of the fundamental signals were injected through the feedback loop of power amplifier. The proposed technique uses the amplifier non-linear characteristic to generate a second third-order IMD signal that is used to cancel the original third-order IMD product at the output. The earlier published results referring to the same novel technique approach are based on the power amplifier simulation for the ideal feedback loop elements. In this paper, the feedback loop components (band pass 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 Libra. Adjusting the phase and amplitude of the loop signals the reduction obtained in third-order intermodulation distortion products are 28 dB in both, for fundamental signals at 2.5 and 2.51 GHz. Simulated results referring to the designed phase shifter and variable attenuator show almost constant characteristics of the relative phase shift and attenuation over frequency range 5-5.2 GHz. Such characteristics provide reduction in third-order IMD3 products of wide spaced carriers.

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