Distortion Improvement of Lossless Feedback Amplifiers Using Augmentation

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Abstract **- A method for improving the distortion characteristics of a class of lossless feedback amplifiers is described, consisting of amplifying and inverting the error voltage at the emitter of the common-base transistor and applying the amplified error voltage to the transistor base. Descriptions are included for active and passive realizations.**

I. Introduction

Lossless feedback amplifiers have long been recognized as a cost-effective means of providing a high degree of dynamic range, in terms of both linearity and noise figure, since their introduction in 1975 [1, 2, 3]. In the ensuing years, various forms have found wide usage in communications receivers [4, 5, 6] and radio astronomy [7, 8, 9]. Their overall economy of design and desirable performance characteristics prompted further study into improvements, and a portion of the results are presented in this paper.

As shown in Fig. 1, the lossless feedback amplifier, consists of a common-base transistor and a feedback transformer. An input voltage V_S enters the amplifier through the input winding of transformer T1, creating an input current to the emitter of transistor Q1, which then causes a collector current to pass through the output winding of transformer T1. By way of the magnetic coupling between the input and output windings of T1, the difference between the input and output currents appears as a correction at the emitter of Q1, thus completing a negative feedback topology. And as a result of the negligible losses involved in the feedback transformer, the amplifier essentially retains the noise figure (NF) of transistor Q1. Signal gain is obtained by virtue of the autotransformer nature of the output winding of T1. The properties of the lossless feedback amplifier of Fig. 1 that are of interest include the power gain, the input impedance, and the collector load resistance which are respectively derived by the relationships [1]:

$$G = M + N + 1 \tag{1}$$

$$R_I = R_L \times \frac{M + N + 1}{M^2} \tag{2}$$

$$R_C = R_L \times (N + M) \tag{3}$$

The collector load resistance is of interest as it represents the limitation of the amplifier with respect to output power compression and saturation, the former of which serves as an indicator of the point at which intermodulation ratio (IMR) expansion occurs for a given transistor and bias condition.

These relationships are largely dependent upon the emitter of transistor Q1 appearing as a virtual ground, which is not necessarily valid but is a convenient assumption in the approximation from which these relationships are derived. Instead, the emitter of Q1 has an input resistance which is approximately described by:

$$R_e = r_e + \frac{r_{bb}}{h_{fe} + 1} \tag{4}$$

where r_{bb} is the base spreading resistance and r_e is the nonlinear incremental emitter resistance of Q1, the latter of which is described by:

$$r_e = \frac{V_E}{I_E} = \frac{V_{BE}}{I_0 \times \varepsilon^{\frac{q V_{BE}}{kT}}} \tag{5}$$

where I_E is the emitter current, I_O is the saturation current, and V_{BE} is the base-emitter voltage of Q1, which has both a quiescent (bias) component and a time-varying (signal) component:

$$V_{BE} = V_{BEQ} + v_{be}(t) \tag{6}$$

Commonly, the emitter resistance is reduced by increasing the transistor bias current, but current design trends dictate that higher degrees of performance be obtained by more power-efficient means, especially with regard to battery-powered field-portable communications equipment. By inspection, the second term of (6) constitutes a signal error voltage, which can be detected and used to provide a correction signal for the common-base transistor, effectively reducing the emitter error voltage and in turn reducing the non-linearitly of the lossless feedback amplifier, the process of which will be referred to herein as augmentation.

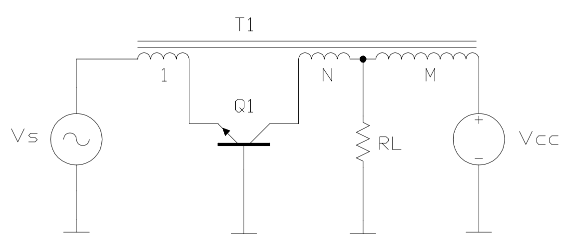

Fig. 1 - Lossless Feedback Amplifier

II. Active Augmentation

One method of providing this correction can be seen in Figure 2, where a common-emitter amplifier is employed to augment the common-base transistor Q1 by amplifying and inverting the emitter signal error voltage and then applying the amplified error voltage to the base [10]. This configuration of transistors was originally proposed by Sidney Darlington in his 1953 patent [11] in which the reduced emitter resistance properties of the configuration were discussed. These properties, however, were never exploited as a means of amplifier linearization, possibly due to the immaturity of transistor circuit design at the time. A similar method has recently been introduced as a means of improving the gain characteristics of cascode amplifiers [12]. In this configuration, the input current is described as:

$$\begin{aligned} I'_E &= I_{E1} + I_{B2} = \\ &= I_{B1}\left(h_{fe1} + 1\right) + \frac{I_{B1}}{h_{fe2}} = \\ &= \left(h_{fe1} + 1 + \frac{1}{h_{fe2}}\right) \times I_{02}\, \varepsilon^{\frac{q V_{BE2}}{kT}} \end{aligned} \tag{7}$$

where h_{fe1} is the signal current gain of transistor Q1, h_{fe2} is the signal current gain of transistor Q2, I_{O2} is the saturation current of Q2, and V_{BE2} is the base-emitter voltage of Q2. Substituting (7) into (5), we find that the apparent emitter resistance becomes approximately:

$$\begin{aligned} r'_e &= \frac{V_E}{I'_E} = \\ &= \frac{V_{BE2}}{\left(h_{fe1} + 1 + \frac{1}{h_{fe2}}\right) \times I_{02}\, \varepsilon^{\frac{q V_{BE2}}{kT}}} \end{aligned} \tag{8}$$

which is a considerable reduction in the nonlinear emitter resistance of the common-base transistor.

III. Passive Augmentation

For applications at high frequencies where NF is a consideration, the use of a common-emitter augmentation amplifier may be impractical as such a circuit may introduce additional noise, thus degrading the NF of the lossless feedback amplifier. For such applications, the use of a simple transformer having a turns ratio of 1:K can give sufficient voltage gain to provide augmentation [10]. Fig. 3 shows such a circuit, where the primary winding of transformer T2 is connected to the emitter of transistor Q1, and the secondary winding is connected in reverse phase to the base. The base voltage and current of transistor Q1 are, respectively:

$$\begin{aligned} V_{BE} &= V_B - V_E = K V_E - V_E = \\ &= -V_E\left(K + 1\right) \end{aligned} \tag{9}$$

$$I_b = \frac{I_E}{h_{fe}} \tag{10}$$

which makes the emitter input current equal to:

$$\begin{aligned} I'_E &= I_E - K I_B = I_E - K\frac{I_E}{h_{fe}} = \\ &= I_E\left(1 - \frac{K}{h_{fe}}\right) \end{aligned} \tag{11}$$

where

$$I_E = I_0 \times \varepsilon^{\frac{q(1+K)\, V_E}{kT}} \tag{12}$$

which allows the apparent emitter resistance to be approximated as:

$$\begin{aligned} r'_e &= \frac{V_E}{I'_E} = \\ &= \frac{V_E}{\left(1 - \frac{K}{h_{fe}}\right) \times I_0\, \varepsilon^{\frac{q V_E (K+1)}{kT}}} \end{aligned} \tag{13}$$

which, compared to (5), shows that the apparent emitter resistance decreases dramatically as the turns ratio of transformer T2 is increased, and is shown in Fig. 4 for a typical BFR92A transistor (h_{fe2} = 90). Notice that with a T2 turns ratio of 1:3 the apparent emitter resistance is reduced to about 5% of that for the unaugmented transistor.

IV. Experimental Results

For the purposes of demonstration, a typical lossless feedback amplifier having the following parameters was constructed

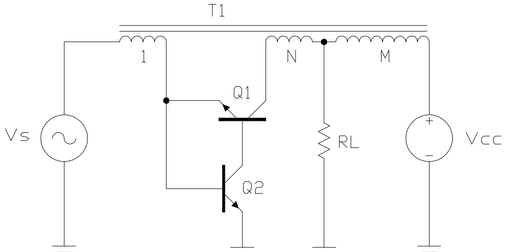

Fig. 2 - Lossless Feedback Amplifier with Active Augmentation

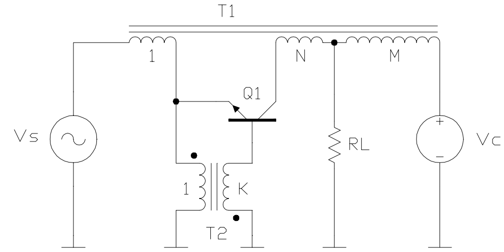

Fig. 3 - Lossless Feedback Amplifier with Passive Augmentation

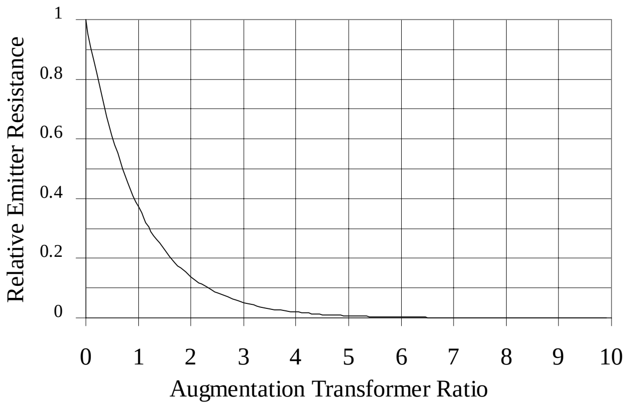

Fig. 4 - Relative Common-Base Emitter Resistance with Passive Augmentation (Typical BFR92A Transistor, h_{fe} = 90)

and evaluated:

Transistor:	Philips BFR92A
	I_C = 10mA
	V_{CE} = 2.65V
	V_{CC} = 5.00V
Transformer:	1:1:2 wound on Fair-Rite 2843-002-402 balun core.
Gain:	+6dB
Impedance:	50 ohms

Gain and intermodulation tests were conducted using signal frequencies of 9.900MHz and 10.100MHz. The results are shown in Fig. 5. Here, it is seen that the IP_3 is +30dBm and that IMR expansion begins moderately at an output power level of +2dBm, approximately 10dB below the 1dB compression point (P_{1dB}).

For the actively augmented lossless feedback amplifier, a second BFR92A device was used for Q2, with a collector current of 2.0mA. Bias conditions for Q1 are the same as for the unaugmented amplifier. It needs to be noted here that a real pole was added to the circuit, consisting of a 100 ohm resistor from the collector of Q2 to the base of Q1 and a 22pF capacitor from the collector of Q2 to ground. This was necessary in order to suppress parasitic oscillations in the region of 600MHz, which is a common problem for lossless feedback amplifiers [9].

The results shown in Fig. 6 indicate that the active augmentation results in a substantial improvement in IP_3 (+39dBm vs. +30dBm) prior to the onset of IMR expansion, but beyond this point the intermodulation (IM) products rise dramatically, actually achieving a level above that for the unaugmented amplifier. Note also that the amplifier gain remains relatively constant for an extended range of output power, resulting in a very abrupt compression knee.

The matter of IMR expansion, a result of the amplifier entering a region of saturation and/or cutoff [13], is a serious design consideration when using feedback amplifiers, more so than with open-loop amplifiers as the nature of the feedback is to drive the amplifier further into the saturation and/or cutoff region(s) as the amplifier is advanced further into compression. Due attention needs to be given to this region of nonlinearity when considering feedback amplifier devices, topologies, and bias conditions.

For the passively augmented lossless feedback amplifiers, a series of three transformers with turns ratios of 1:1, 1:2, and 1:3 were constructed using a Fair-Rite 2843-002-402 balun core. Bias conditions for Q1 are the same as for the two prior examples.

Fig. 7 is a typical example of the results obtained for the passively augmented lossless feedback amplifiers, in this case using a 1:2 augmentation transformer. Here, the IM products do not increase as dramatically as with the actively augmented circuit at the onset of IMR expansion, and the level approaches that of the unaugmented amplifier as the amplifier is driven further into compression. The IP_3 for this circuit is a respectable +37dBm, and this improvement in performance comes without any increase in power supply consumption, which is an attractive design consideration. In addition, the augmentation transformer does not introduce any additional noise sources, but instead places the amplifying transistor within a virtually lossless embedding topology.

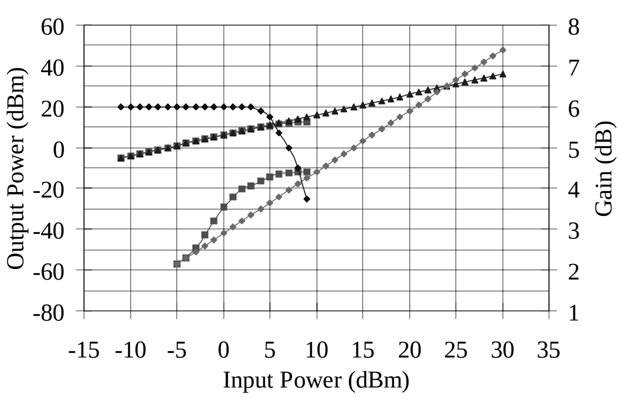

Fig. 5 - Lossless Feedback Amplifier Intermodulation Performance (BFR92A, V_{CE} = 2.65V, I_C = 10mA)

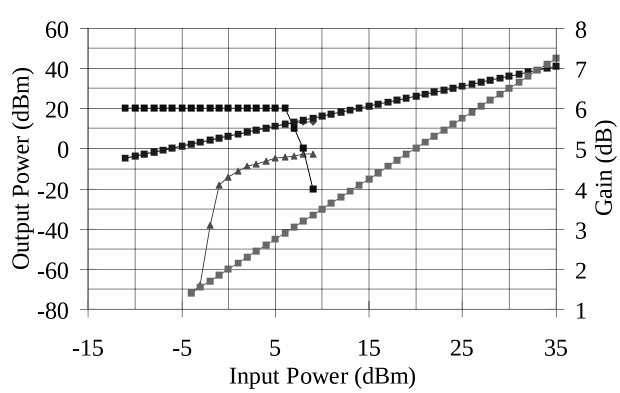

Fig. 6 - Lossless Feedback Amplifier Intermodulation Performance (Active Augmentation)

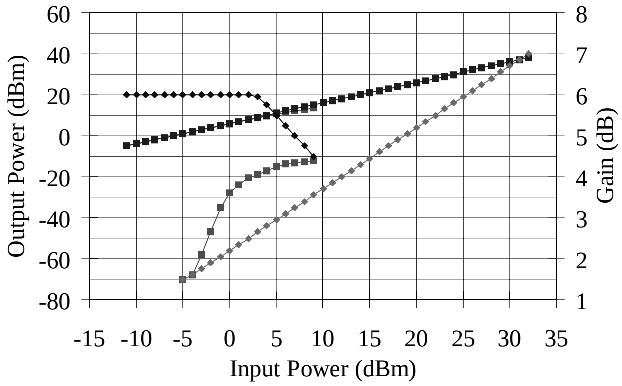

Fig. 7 - Lossless Feedback Amplifier Intermodulation Performance (1:2 Passive Augmentation)

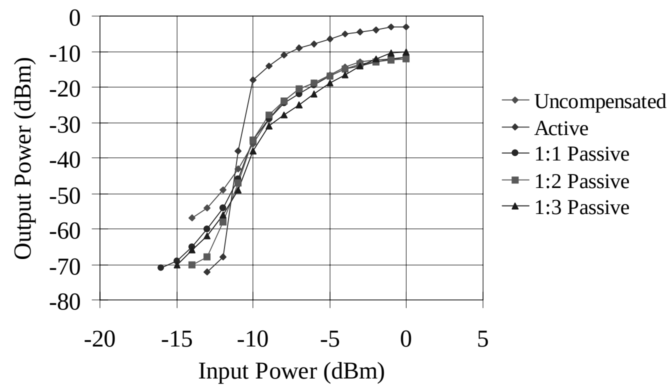

Fig. 7 - Augmented Lossless Feedback Amplifier Intermodulation Performance

V. Performance Comparison

A comparison of the IM performance of the circuits described here is shown in Fig. 7. Note that the onset of IMR expansion is fairly consistent at about +2dBm signal output power (see Fig. 5 and discussion). It is easily noticed here how the IM products for the actively augmented circuit exceed those for the unaugmented circuit above the onset of IMR expansion, whereas the passively augmented circuits generally conform to the performance of the unaugmented circuit in this region. For most applications, the use of passive augmentation should be sufficient.

Table 1 lists the IP_3 and P_{1dB} performance for the described circuits, and here it can be readily seen that the actively augmented circuit gives slightly better intermodulation performance than the passive circuits, but when the IMR expansion characteristics shown in Fig. 7 are taken into account it is seen that the passively augmented circuits have a definite overall advantage. Also, the active augmentation gives very little advantage with respect to the compression point, and the 1:3 passive realization gives comparable results in this area.

VI. Conclusions

The intermodulation performance of single-transformer lossless feedback amplifiers can be improved below the onset of intermodulation expansion by the application of augmentation, wherein the transistor emitter voltage is considered as an error signal, which is then amplified, inverted, and applied to the transistor base. The process of augmentation can be accomplished either actively, using a common-emitter amplifier, or passively, using a two-winding transformer, the result being an apparent reduction in the common-base transistor emitter impedance, a primary source of nonlinearity.

TABLE 1
Augmented Lossless Feedback Amplifier Intermodulation Performance

Configuration	IP_3	P_{1db}
Unaugmented	+30dBm	+12dBm
Active	+39dBm	+12.5dBm
Passive 1:1	+34dBm	+12dBm
Passive 1:2	+37dBm	+12dBm
Passive 1:3	+35dBm	+12.5dBm

References

[1] Norton, David E. and Allen F. Podell, “Transistor Amplifier with Impedance Matching Transformer,” U.S. Patent 3,891,934, 24 June 1975.

[2] Norton, David E., “High Dynamic Range Transistor Amplifiers Using Lossless Feedback,” *Proceedings of the 1975 IEEE International Symposium on Circuits and Systems*, pp. 438-440.

[3] Norton, David E., “High Dynamic Range Transistor Amplifiers with Lossless Feedback,” *Microwave Journal*, May 1976, pp. 53-57.

[4] Golovin, O.V. and I.K. Agilar, "Methods of Improving Linearity in Professional SW Receivers," *Telecommunications and Radio Engineering*, Vol. 39, No. 1, Jan 1984, pp. 69-74.

[5] Zhang, Y. and Y. Peng, "Design and Analysis of Noiseless Feedback Amplifiers Using S Parameters," *Microwave Journal*, Feb 1998, pp. 112-124.

[6] Makhinson, J., "A High-Dynamic-Range MF/HF Receiver Front End," *QST*, Feb 1993, pp. 23-28.

[7] Abranin, E.P. and Y.M. Bruk, “Highly Linear Broadband Amplifiers with Feedback,” *Telecommunications and Radio Engineering*, Vol. 41/42, No. 5, May 1987, pp. 90-91.

[8] Abranin, E.P. and Yu. M. Bruck, “Highly Linear Broadband Amplifiers with Lossless Feedback, Part 1: Theory,” *International Journal of Electronics*, 1990, Vol. 68, No. 5, pp. 743-756.

[9] Abranin, E.P. Yu. M. Bruck, and A.A. Konovalenko, “Highly Linear Broadband Amplifiers with Lossless Feedback, Part 2: Experience of Development,” *International Journal of Electronics*, 1990, Vol. 69, No. 3, pp. 345-357.

[10] Patent pending.

[11] Darlington, Sidney, “Semiconductor Signal Translating Device,” U.S. Patent 2,663,806, 22 December 1953.

[12] Sackinger, Eduard and Walter Guggenbuhl, "A High-Swing, High Impedance MOS Cascode Circuit," *IEEE Journal of Solid-State Circuits*, Vol. 25, No. 1, February 1990, pp. 289-297.

[13] Trask, C., "Nonlinear Distortion in Bipolar Transistors," *QEX*, November/December 1998, pp. 37-42.