Lossless Feedback Amplifiers:

Theory and Advanced Techniques

by

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Introduction

Among the topics of interest to designers of telecommunications and radio equipment, intermodulation (IMD) and noise figure (NF) are of particular interest as they are limiting factors in the performance of radio receivers. The minimum detectable signal (MDS) of a receiver is determined by way of the NF, and together with the third-order intercept point (IIP3) determines the spurious-free dynamic range, commonly referred to as dynamic range.

The earliest stages of a receiver determine the overall dynamic range, and it is important that close attention be given to the design of the low-noise amplifier (LNA) stage(s) immediately after the antenna input terminals. The NF is very much a matter of proper device selection together with proper device impedance matching and sufficient decoupling of power supply noise.

The IMD performance of an amplifier is very dependent upon the biasing conditions, as well as the linearity of the device itself. Unfortunately, the bias conditions that provide high IMD performance very often conflict with otherwise excellent NF performance, therefore a judicious compromise needs to be considered in the architecture of the receiver system design so that the LNA design remains practical yet near optimal in terms of overall gain, NF, IMD, and bandwidth performance objectives.

Series/Shunt Feedback Amplifiers

One method to obtain good IMD performance is to make use of negative feedback in the amplifier design. For RF amplifiers, two methods of negative feedback that find wide application are series/shunt feedback and lossless feedback.

The series/shunt feedback amplifier topology, shown in Fig. 1, was devised and patented by Leonard Seader and James Sterett of Avantek (1). Shown in basic form in Fig. 2, the amplifier consists of a transistor and two resistors. The design equations for the two resistors are quite simple:

$$R_{IN} = R_{OUT} = \sqrt{R_{FB} \times R_E} \tag{1}$$

$$\frac{v_{out}}{v_{in}} = \frac{i_{out}}{i_{in}} = \frac{1-\sqrt{R_{FB}/R_E}}{2} \tag{2}$$

from which we now define the amplifier voltage gain A_V as:

$$A_V = \left|\frac{v_{out}}{v_{in}}\right| \tag{3}$$

and the amplifier power gain is now:

$$G = 20\,log\;A_V \tag{4}$$

and we can now derive the values of the two resistors as:

$$R_{FB} = R_L\left(2\,A_V + 1\right) \tag{5}$$

$$R_E = \frac{R_L^2}{R_{FB}} \tag{6}$$

Although the series/shunt amplifier may be convenient, the use of resistors, whether carbon or metal film, is deterimental to otherwise good NF performance, and the overall NF of the series/shunt amplifier is often many dB beyond the NF of the transistor.

The series/shunt amplifier can provide a

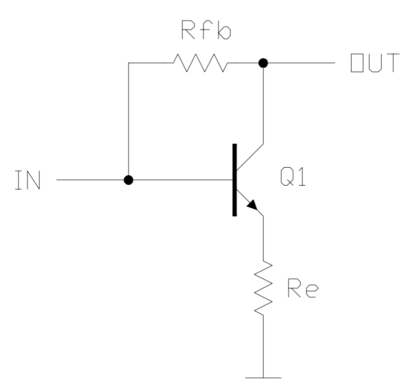

Figure 1 - Series/Shunt Feedback Amplifier

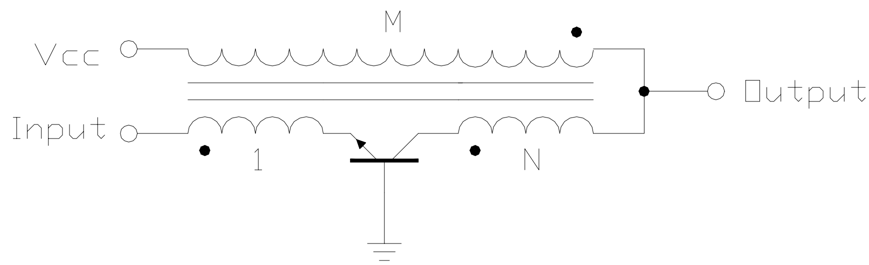

Figure 2 - Lossless Feedback Amplifier

good level of saturated power since the collector is coupled directly to the load. However, the degree of linearization is dependent upon the difference between the signal gain of the transistor and the closed loop gain. As a general rule of thumb, the transistor signal gain should be 3dB or more greater than the closed loop amplifier gain.

An interesting innovation in this topology is to use a cascode pair in place of the single transistor. Although this configuration decreases the amplifier saturable power, it does increase the high cutoff frequency.

Lossless Feedback Amplifiers

Perhaps the single most significant development in high dynamic range amplifiers has been that of the lossless feedback amplifier. Conceived and patented by David Norton and Allen Podell of Adams-Russell (2), this topology is often referred to as a Norton amplifier, and sometimes as noiseless feedback.

As shown in Fig. 2, the amplifier consists of a transistor and a three-winding transformer. Overall, the performance is less dependent on the transistor and more dependent on the transformer, particularly the coupling coefficient between the three windings. Neglecting the finite emitter input resistance and the induced losses and less than unity coupling of the transformer, the input impedance of the amplifier is simply:

$$R_{IN} = R_L \frac{M + N + 1}{M^2} \tag{7}$$

and the amplifier gain is:

$$G = 20 \log M \tag{8}$$

For an amplifier whose input impedance is equal to the load impedance, the turns ratios of M and N are related by:

$$N = M^2 - M - 1 \tag{9}$$

An important factor to take into consideration is the collector load resistance:

$$R_C = R_L (M + N) \tag{10}$$

This collector load resistance limits both the saturable power and the high frequency cutoff of the amplifier, and is rarely given much attention. In Table 1 below, it can be readily seen that as the turns ratio of the transformer are increased to obtained higher gain, R_C rises rapidly, depriving the amplifier of both dynamic range and bandwidth, as we shall see shortly.

With respect to the saturable power, the maximum peak power that can be obtained is

N	M	Rc Ohms	Gain dB
1	2	150	6.02
5	3	400	9.54
11	4	750	12.04
19	5	1200	13.98

Table 1 - Lossless Feedback Amplifier Transformer Ratios

a result of R_C and the combination of the quiescent collector-emitter voltage (V_{CE}) and the collector saturation voltage ($V_{CE\ SAT}$) of the transistor:

$$P_{max} = \frac{(V_{CE} - V_{CE\,SAT})^2}{R_C} \qquad (11)$$

and the average power, which is what we typically measure, is 3dB less than this.

The high cutoff frequency dependency on R_C is a result of the collector-emitter and collector-base capacitances of the transistor (C_{oss} and C_{rss}, respectively) as well as the net capacitance across the output windings of the transformer (C_{NM}):

$$f_{max} = \frac{1}{R_C(C_{oss} + C_{rss} + C_{NM})} \qquad (12)$$

Other factors, such as the leakage inductances of the transformer and stray capacitance to ground, serve to further lower f_{MAX}.

An additonal factor that limits the overall performance of the lossless feedback amplifier has to do with the approximation that Norton and Podell made in formulating the design equations, which was the assumption that the emitter input resistance (R_E) of the transistor is negligible. This is, of course, very convenient in both the overall theory and the design process, but in reality R_E is not only finite but it is also nonlinear, and both of these qualities detract from the overall potential linearity of the amplifier.

We could, or course, increase the transistor bias current so as to reduce R_E, but this would have other impacts on the amplifier. First, it would increase the supply current and therefore reduce the amplifier power efficiency. In addition, transistors typically have their best NF and linearity characteristics at lower collector currents,

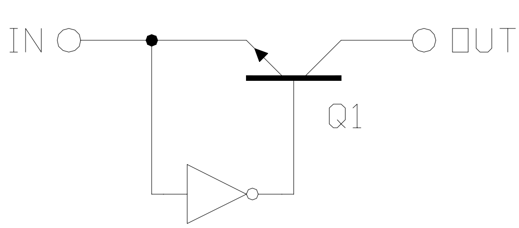

Figure 3 - Basic Linearity Augmentation

Linearity Augmentation

Since the issuance of the initial patent, little has been done to continue the further development of the lossless feedback amplifier, which is a bit surprising given the emphasis on raising the signal-to-noise ratio (SNR) and IMD performance of telecommunications equipment so as to aedquately get above the noise floor and further improve the overall dynamic range. To this effect, linearity augmentation was devised and patented as a means of overcoming the intrinsic finite and nonlinear R_E that compromises the performance of common-base (3, 4) and lossless feedback (5, 6) amplifiers, as well as to improve the NF. Basically, as shown in Fig. 3, the signal voltage at the emitter of the transistor is sensed by an inverting amplifier and the output is applied to the base. For a simple voltage amplifier having a voltage gain of A_V, the reduction of R_E is:

$$R'_E = \frac{R_E}{A_V + 1} \qquad (13)$$

where the emitter resistance R_E is determined by the familiar relationship:

$$R_E = \frac{V_{BE}}{I_0\, \varepsilon^{\frac{q V_{BE}}{k T}}} \qquad (14)$$

Since wideband voltage amplifiers hav-

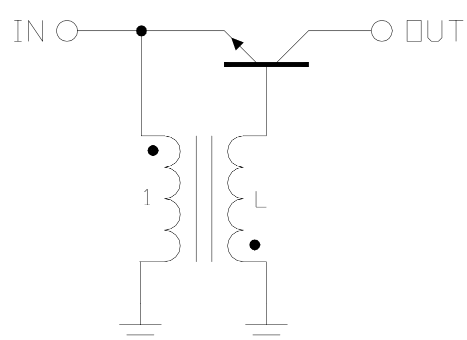

Figure 4 - Passive Augmentation

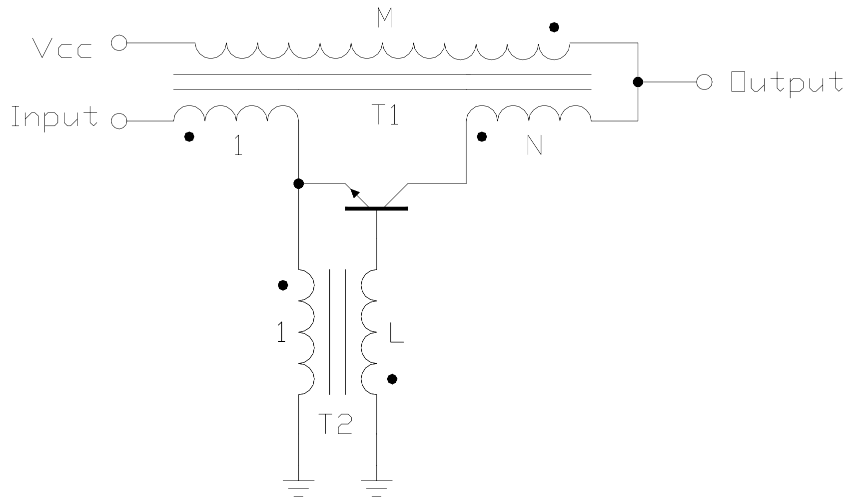

Figure 5 - Lossless Feedback Amplifier with Passive Augmentation

ing simultaneously high gain and low noise at RF frequencies are somewhat impractical to achieve, two alternative methods were devised, being passive and active augmentation. With passive augmentation, shown in Fig. 4, the voltage amplifier is replaced with a simple two-winding transformer having a windings ratio of L, where the value of L is arbitrary. With this method, the reduction of R_E becomes approximately:

$$R'_E = \frac{R_E}{\left(1 - \frac{L}{h_{fe}}\right)(L+1)} \tag{15}$$

For transistors having h_{fe} of 100, a reduction of R_E in the order of 95% is realizable with a transformer having a turns ratio of 1:3. Fig. 5 illustrates a lossless feedback amplifier with passive augmentation.

Passive augmentation also has benefits in terms of NF. Using Fig. 6 as a reference, the noise of the transistor (Q1) passes through the primary winding ot the transformer T2, where it is amplified and inverted and then coupled to the base of Q1, resulting in a reduction of noise by approximately 1/L. The noise source v_{th} represents the noise added by the bulk and induced losses as well as the Barkausen noise of the core of T2. Due to the nature of trans-

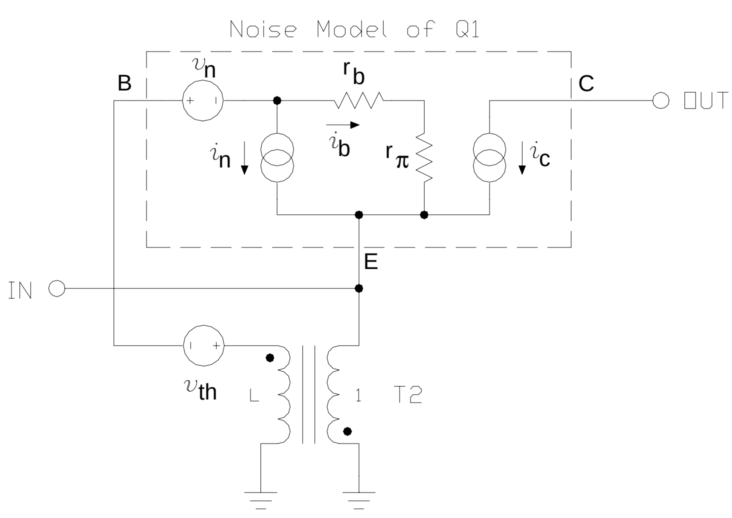

Figure 6 - Noise Model of Passive Augmentation

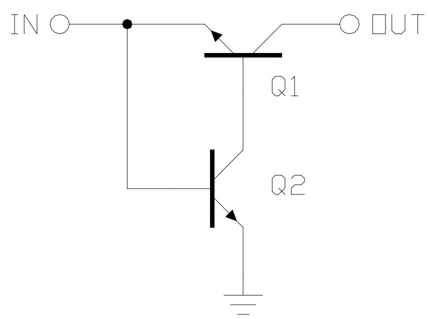

Figure 7 - Active Augmentation

formers, this reduction is limited to the thermal, or Nyquist noise of the transistor as the 1/f, or "flicker" noise will most likely be below the low cutoff frequency of the transformer.

There are, of course, practical limitations to the windings ratio of the augmentation transformer. As the ratio is increased, parasitics such as leakage inductances and both intra- and interwinding capacitances can limit the frequency bandwidth. To further improve the reduction of R_E, active augmentation, as shown in Fig. 7, can be employed. Here, the second transistor Q2 amplifies and inverts the signal voltage at the emitter of the first transistor Q1, resulting in a substantial reduction in R_E:

$$R'_E = \frac{R_{E2}}{\left(h_{fe1} + 1 + \frac{1}{h_{fe2}}\right)} \qquad (16)$$

where R_{E2} is the emitter resistance of transistor Q2 and h_{fe1} and h_{fe2} are the signal current gains of transistors Q1 and Q2, respectively. A schematic depicting the application of augmentation in a lossless feedback amplifier is shown in Fig. 8.

Provided that Q2 is properly biased and matched for it's best noise performance, the use of active augmentation provides a substantial reduction of the 1/f noise of the amplifier transistor Q1, but at the same time will add some amount of thermal noise above that which would be added by use of passive augmentation. Shown in detail in the noise model of Fig. 9, the resistor R_{B1} is the biasing resistor (or resistors) shared by the collector of Q2 and the base of Q1. The noise source v_{ps} represents the noise from the power supply. Resistor R_{E1} is the emitter biasing resistor of Q1, and resistor R_{B2} is the base resistor added between R_{E1} and the base of Q2 to properly match Q2 for its best noise performance. The noise source v_{th} represents the thermal noise added by the three resistors.

An added benefit of active augmentation is the reduction of even-ordered IMD products, such as 1x1, 2x2, 3x3, etc. that exist at low or baseband frequencies. The transformer used with passive augmentation might not conduct these signals as they would very likely fall below the low cutoff frequency of the transformer. However, in an amplifier using direct-connected active augmentation these signals would be corrected down to DC.

Both passive and active augmentation have benefits and shortcomings. Passive aug-

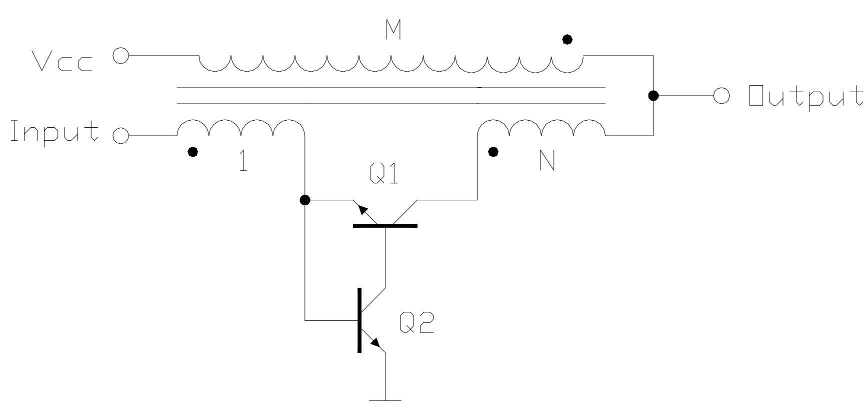

Figure 8 - Lossless Feedback Amplifier with Active Augmentation

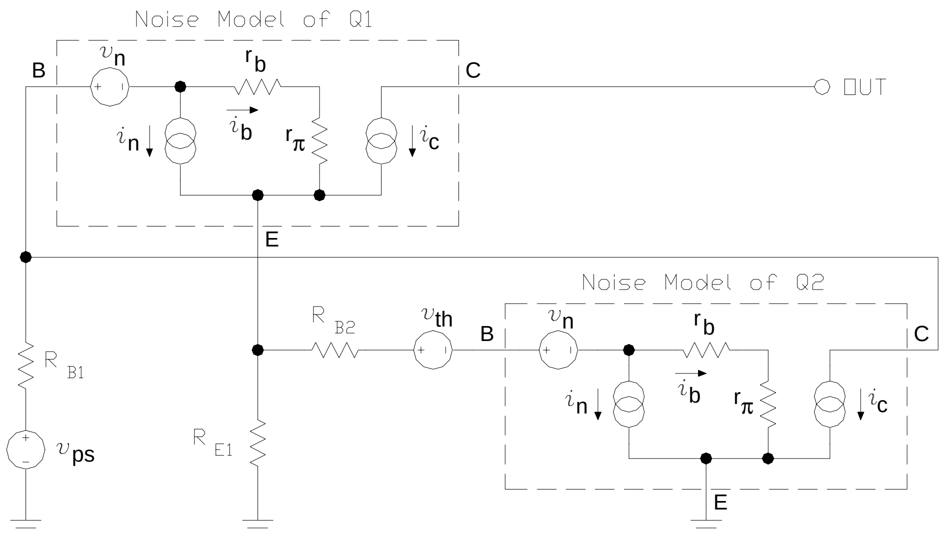

Figure 9 - Noise Model of Active Augmentation

mentation provides the needed reduction in the emitter resistance and some reduction in the thermal noise, but due to the nature of transformers it has little effect on the even-order baseband IMD products or the 1/f noise. Active augmentation provides a higher degree of emitter resistance reduction plus effective reduction in 1/f noise and basebnd IMD products, but at the same time has the potential to add thermal noise.

The two methods may be combined so that their benefits may complement each other, thereby overcoming each others' shortcomings. This method is known as tandem augmentation, and is illustrated in the schematic of Fig. 10. Here, the transformer T2 provides the inband passive augmentation to reduce inband IMD products along with some reduction in the thermal components contributing to the NF, while the transistor Q2 provides the baseband augmentation to correct the baseband IMD products as well as the 1/f components of the NF. The bypass capacitors at the base and collector of Q2, together with transformer T2 create a crossover network that determines the

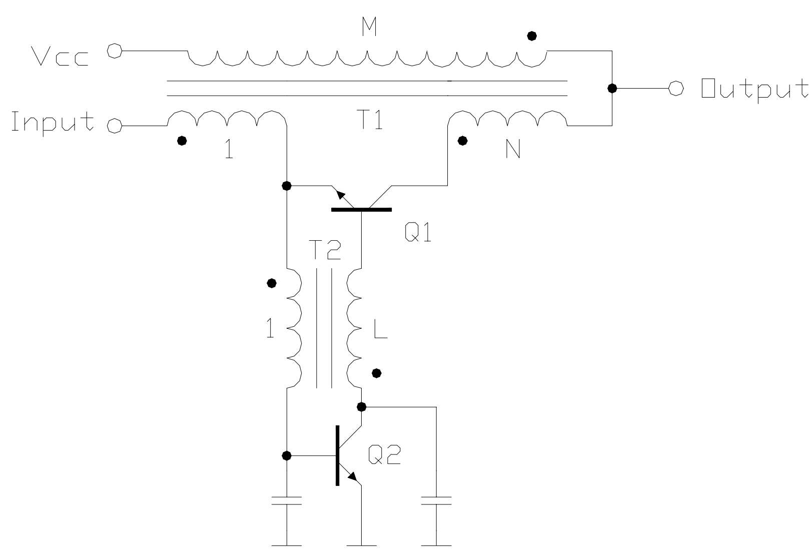

Figure 10 - Lossless Feedback Amplifier with Tandem Augmentation

point at which the passive and active augmentation dominate.

High Frequency Performance

It was mentioned earlier that the collector load resistance R_C is a significant factor in the high frequency limit of the lossless feedback amplifier. From Eq. 12, it is obvious that there is little that can be done to substantially reduce the various capacitances, and reducing R_C requires that the turns ratio of the feedback transformer be reduced, which will result in an amplifier of lower gain.

An alternative is to provide some form of current gain within the amplifier, and this can be done easily by adding an autotransformer to the input of the common-base amplifier transistor. Such a method is shown in Fig. 11, where a tap has been added to the emitter side of a passive augmentation transformer, resulting in what is called compound passive augmentation. Here, the common-base amplifier has a current gain of K+1, and the base side of the transformer performs in the same manner as with simple passive augmentation.

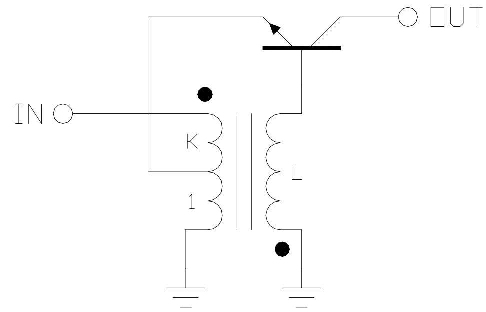

Figure 11 - Common Base Amplifier with Compound Passive Augmentation

The implementation of compound passive augmentation in a lossless feedback amplifier is illustrated in the schematic of Fig. 12. The voltage gain of the amplifier is still equal to M, as before, and the gain remains as:

$$G = 20 \log M \tag{17}$$

Now, the input resistance has become:

$$R_{IN} = R_L \frac{(K+1)(M+N)+1}{M^2} \tag{18}$$

For an amplifier whose input impedance is equal to the load impedance, the turns ratios of M and N are related by:

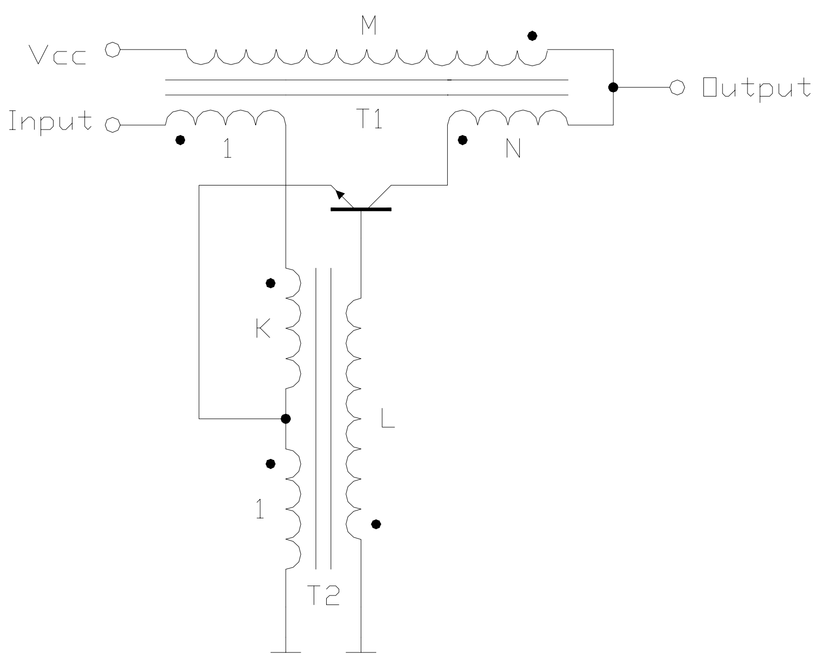

Figure 12 - Lossless Feedback Amplifier with Compound Passive Augmentation

$$N = \frac{M^2 - 1}{K+1} - M \quad (19)$$

and the collector load resistance now becomes:

$$R_C = R_L \frac{M+N}{K+1} \quad (20)$$

which shows that by compound passive augmentation can substantially reduce R_C with little if any added circuit complexity. Table 2 lists a number of combinations that can be realized to provide a significant increase in amplifier gain while retaining or even improving the high cutoff frequency by way of a reduction in R_C.

K	N	M	Rc Ohms	Gain dB
1	1	3	100	9.54
1	7	5	300	13.98
2	1	4	83	12.04
2	3	5	133	13.98
2	9	7	267	16.90
3	1	5	75	13.98
3	5	7	150	16.90
3	11	9	250	19.08

Table 2 - Compound Lossless Feedback Amplifier Transformer Ratios

Tandem augmentation may also be applied to the compound augmented lossless feedback amplifier of Fig. 12, as illustrated in the schematic of Fig. 13. A number of these amplifiers have been constructed and tested, and with common parts such as a 2N2222 for Q1 and an MPSA14 Darlington pair for Q2, an OIP3 of +45dBm and a NF approaching 1dB is easily achieved.

The techniques described herein extend the Norton patent into a strongly enhanced and optimized design in which NF, IMD, gain, and high frequency performance can be improved without compromising power efficiency and circuit complexity, objectives which were difficult to achieve with earlier designs. Furthermore, these techniques demonstate that such amplifiers can be built with commonly obtainable parts, making the improved performance economically attractive. It is hoped that others will be encouraged to extend or develop new technologies in this important area of LNA design.

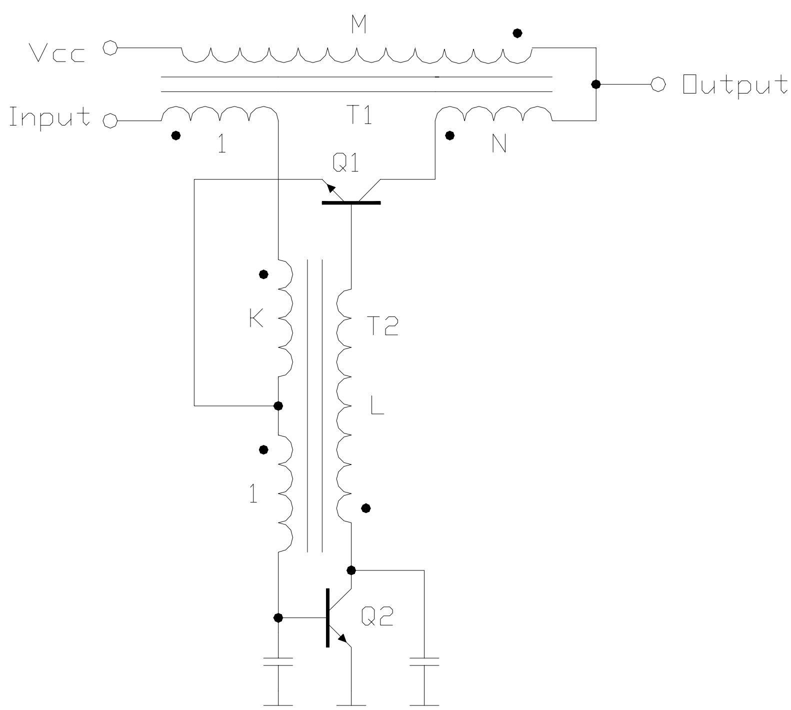

Figure 13 - Lossless Feedback Amplifierwith Compound Tandem Augmentation

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