# Common Base Transformer Feedback Norton Amplifiers

### *Formerly Ultralinear 2N5109 And 2N3053 Amplifiers*

### Dallas Lankford, 8 VI 94, rev. 21 V 07

For about a year I have been experimenting with extremely high intercept amplifiers using 2N5109 bipolar junction transistors (BJTs). These kinds of amplifiers, known as common base transformer feedback (CBTF) amplifiers, and also known as common base noiseless feedback amplifiers, if properly designed and used, offer MW DXers (and SW DXers, though their needs are not as extreme as MW DXers) amplifiers with extremely low levels of 2nd and 3rd order intermodulation distortion (IMD2 and IMD3), which, in turn, offer much higher levels of strong signal handling performance than have been available previously. CBTF amps can provide greatly improved performance for balanced two foot air core loop amps, for low-gain tuned preselector amps, for broadband phasing system amps, and for other similar applications where extremely high 2nd and 3rd order input intercepts (IIP2 and IIP3) are needed. Appropriately configured, a single BJT CBTF amp can have IIP3 greater than +35 dBm throughout the MW band and IIP2 greater than +46 dBm throughout the MW band. With a pair of BJT CBTF amps configured push-pull, IIP2 greater than +95 dBm can be achieved throughout the MW band. The purpose of this note is to summarize my experiences with CBTF amps using 2N5109's, and to provide enough information for other DXers to construct and use these kinds of amplifiers.

CBTF amps were first described by Dr. David E. Norton in his pioneering May 1975 **Microwave Journal** article, "High dynamic range transistor amplifiers using loss less feedback." The amplifiers were patented (U.S. Patent Nos. 3,426,298;1969, and 3,624,536; 1971, and 3,891,934; 1975), which, perhaps, explains why CBTF amps have not been widely used in the past. According to reliable sources, these patents have expired. However, if you decide to produce and sell such amps, you should obtain legal counsel to verify this information.

In that article, only transformer turns ratios and corresponding amp gains were given; no information about DC power and biasing was given; see Fig. 1 at left below. A theoretical circuit analysis based on simplifying
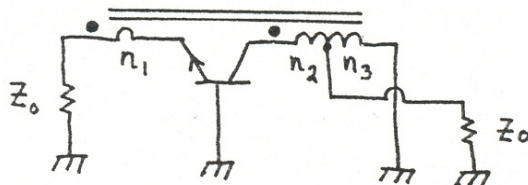


Fig. 1

assumptions was given as follows. Assuming that a common base BJT amp has zero input impedance, and infinite output impedance (neither of which is true), and assuming unity current gain (which is approximate true provided the operating frequency of the amp is well below the cutoff frequency of the BJT), a two way impedance match to $Z_0$ will be obtained if the transformer turns ratios $n_1{:}n_2{:}n_3$ satisfy the turns ratio condition l:n:m where $n = m^2 - m - 1$. Permissible ratios include 1:1:2, 1:5:3, 1:11:4 1:19:5, and so on. Also with the above simplifying assumptions, power gain is $m^2$. Thus, a 1:1:2 amp has gain 4, or 10 1og(4) = 6.02 dB, a 1:5:3 amp has gain 9, or 9.54 dB, a 1:11:4 amp has gain 16, or 12.04 dB, a 1:19:5 amp has gain 25, or 13.98 dB, and so on. The number of turns of wire on the transformer may be varied to adjust frequency range of the amp as long as the appropriate ratios are maintained. For example, a 1:5:3 amp may have 1:5:3, or 2:10:6, or 3:15:9 turns for $n_1{:}n_2{:}n_3$ and so on.

The phasing dots of the transformer in Fig. 1 should be observed. A BJT CBTF amp uses negative feedback, so that reversal of the phasing of the feedback link $n_l$ would provide positive feedback, which would likely cause the amp to oscillate, and in any case would change amp gain, and degrade IMD performance and two way impedance match to $Z_0$.

Although it was not explained in Norton's article, the meaning of the expression "a two way impedance match to $Z_0$" is, apparently, that if one of these BJT CBTF amps works into a load of $Z_0$ ohms real, then the input

impedance of the amp is $Z_0$ ohms. In other words, the input impedance of a BJT CBTF amp is dependent on the load impedance; namely, the input impedance is equal to the load impedance. But there is more to it than that. The transformer of the BJT CBTF amp is a broadband transformer, with a frequency range which depends on transformer parameters. So the frequency range for which a BJT CBTF amp provides a two way match to $Z_0$ is, presumably, no greater than the frequency range of the transformer. Also, the frequency range of a particular transformer depends on the source and load impedances of the transformer. This limits the range of values of $Z_0$ for which a particular transformer will provide a two way impedance match to $Z_0$ over a given frequency range. Based on experiments and measurements with a number of transformers, I have found that if the usual principles of broadband transformer design are adhered to, then a BJT CBTF amp using 2N5109's has about the same frequency range as the transformer alone for a given $Z_0$. For many applications, $Z_0 = 50$ ohms, so the usual transformer design for 50 ohms should be used.

The two way impedance match to $Z_0$ is an ideal characteristic of BJT CBTF amps based on a mathematical derivation which uses simplifying assumptions that are not true for actual common base amps, as pointed out above. Based on measurements with several BJT CBTF amps, I have found that if such an amp is terminated in a load of 50 ohms real, then the input impedance tends to be about 60% to 80% of 50 ohms. I have not attempted to determine whether a perfect match to input impedance (using a broadband matching transformer) would increase the IIP2 and IIP3 of these BJT CBTF amps because the intercepts are already so high that there seems to be no need to raise them slightly higher by this means. Also, there are other easier means to raise the intercepts higher. For example, a 1:11:4 CBTF amp with appropriate transformer and biasing adjusted for 20 mA collector current and with appropriate bypassing and coupling capacitors has flat gain of 12 dB nominal (about 10.8 dB) from about 100 kHz to beyond 30 MHz, while the IIP3 is about +38 dBm from 10 MHz to 30 MHz, but falls off slowly below 10 MHz to about +34 dBm at about 1.6 MHz, and to about +27 dBm at about 455 kHz. This decrease of IIP3 as frequency decreases seems to be normal and due to the diode junctions of BJT's. For higher IIP3 within and below the MW band, one may use a 2:11:4 transformer, which has a flat gain of about 6 dB, and IIP3 greater than +35 dBm for all frequencies greater than 455 kHz. The two way impedance match of a 2:11:4 transformer is not perfect either, and gives an input impedance of about 160 % of $Z_0$. For a 2:11:4 CBTF amp working into a 50 ohms real load, this would be an input impedance of about 80 ohms, which is still a reasonably good match to a 50 ohms source impedance. Not also (above) that the gain formula (theoretical gain) does not accurately describe measured gains for higher gains. Actual gains are generally less than theoretical gains at higher gain levels.

The example of a 2:11:4 CBTF amp above illustrates three of the general principles of negative feedback, namely that (l) as negative feedback is increased, power gain is decreased, (2) as negative feedback is increased, linearity improves (IIP3 increases), and (3) as negative feedback is increased, input impedance of the feedback amp is increased. An excellent discussion of negative feedback amplifiers is contained in Chapter 17 of the book *Electronic Devices And Circuits*, by J. Millman and C. Halkias, McGraw-Hill Book Co., 1967. There a fourth general characteristic of negative feedback amplifiers is discussed, namely the tendency of amp noise to decrease as negative feedback is increased, but I have not observed this characteristic in any of my experiments and measurements, perhaps because noise from other sources obscured any noise reduction which resulted from the higher feedback 2:11:4 CBTF amp.

My experiences with 2:11:4 CBTF amps, as related above, suggest that a fertile area for future investigation may be to study the performance characteristics of CBTF amps with transformer turns ratios other than the ideal ratios (1:1:2, 1:5:3, 1:11:4, and so on) originally proposed by Norton (and repeated by other writers). For example, it appears that the turns ratios for a near-perfect two way impedance match to $Z_0$ have yet to be determined for 2N5109's, and since such amps have yet to be studied, their power gains and their input intercept characteristics are as yet unknown. And it appears that the only way to develop such amps, if they can be developed, is by trial and error, i.e., by winding transformers with different turns ratios and measuring the closeness of two way impedance match to $Z_0$, measuring amp gain, measuring IIP2 and IIP3 for various collector currents, and so on. One could also attempt to develop a more accurate mathematical model of CBTF amps beginning with the usual two-port hybrid models, but the complexity of such a project appears to be considerable.

Relatively little information appears to have been published regarding biasing CBTF amps using 2N5109's. The November 1984 **Ham Radio** column, "VHF/UHF World," by Joe Reisert, W1JR, contains the only biasing information I have found for a 2N5109. The same biasing was also suggested for an NEC NE4163B transistor. Another biasing arrangement, one for Motorola MRF586 BJT's, requiring +6 VDC and -6 VDC power sources, was discussed by J. Makhinson in his Feb. 1993 **QST** article,"A high-dynamic-range MF/HF receiver front end." However, the dual polarity power requirement makes Makhinson's approach difficult to implement, and based on experiments I have done, there is no improvement in linearity due to the dual polarity power arrangement. In part 2 of his Dec. 1981 **Ham Radio** article, "Communications receivers for the year 2000," Dr. Ulrich Rohde briefly summarized what Dr. David Norton had already published in 1976, and gave an example (in his Fig. 7) of an elaborate two-stage amplifier using the noiseless feedback concept. The two-stage amp used Siemens BFT66 BJT's, which are not widely available in the U.S.A. Due to the complexity of that two-stage amp, and because of the difficulty of obtaining BFT66 BJT's, it does not seem appropriate for hobbyist or consumer grade applications at MW's and SW's. Another example of CBTF amps was given in Rohde's Nov. 1992 **QST** article, "Recent advances in shortwave receiver design," namely, the use of an AGC controlled BFT66 CBTF IF amp (Fig. 11 of his article) in a Rohde & Schwarz EK-890 communications receiver.

Two schematics, one for the basic CBTF amp, and the other for a push-pull amp using two matched CBTF amps, are given below in Fig. 2. For MW band use, C1, C2, C3, C4, and C5 should be 0.1 or 0.2 uF, RFC1 should be 1 mH, RFC2 may be 100 uH or greater, up to 1 mH, FB should be an Amidon ferrite bead, type FB-101-64, T1 should be a Amidon FT-50-75 ferrite toroid core with $n_1 = 1$ turn, $n_2 = 11$ turns, and $n_3 = 4$ turns #24 enameled copper wire, and the 75 material core should be wrapped with thick Teflon tape before winding the wire turns, where the wire turns are spaced evenly around the entire circumference of the toroid core, R2 = 1000 ohms, R3 = 4700 ohms, and R3 = 10 ohms for +V DC = +9 volts DC, R1 should be a 100 ohm adjustable pot (I like Spectrol 25 turn 1/2 watt cermet top adjust pots in series with a 10 ohm 1/4 watt fixed resistor, the latter to prevent frying Q1 if the pot is accidentally adjusted to zero ohms), and Q1 is a 2N5109. R1 is adjusted for whatever collector current is desired (up to about 16 mA if Q1 is not heat sinked, and up to about 30 mA if Q1 is heat sinked with a Mouser 567-7-120-BA heat sink rated at 35 degrees C/W ther. res.). The current drain (setting of R1) determines the amp IIP3 and IIP2. More information about the relationship between current drain and IIP3 and the relationship between current drain and IIP2 will be given below.



Fig. 2

Basic Amp
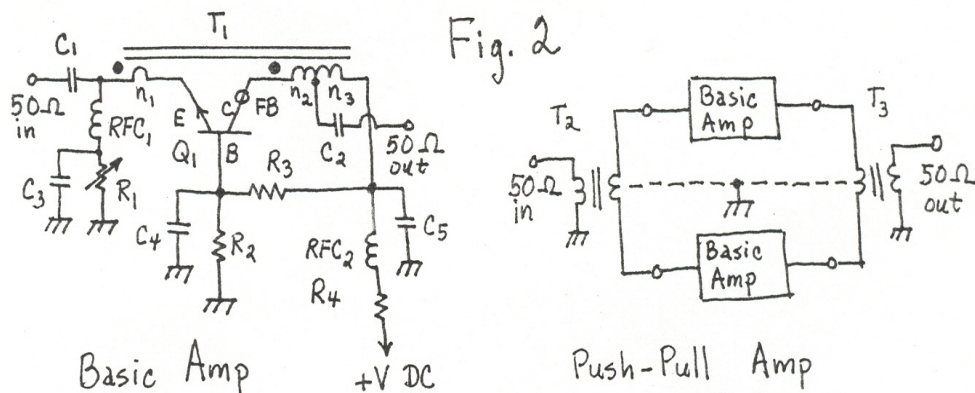
Push-Pull Amp

A push-pull (sometimes called balanced) CBTF amp requires a pair of basic amps configured as shown above in Fig. 2. For a push-pull amp with a frequency range of about 100 kHz to beyond 30 MHz, T2 and T3 should be Amidon FT-50-75 ferrite toroid cores, wrapped with thick Teflon tape, and wound with 8 bifilar turns of #24 enameled copper wire. The basic amps should be shielded from each other as shown in Fig. 2; otherwise, interaction between the individual amps may cause instability or degrade the extremely high IIP2 which this kind of push-pull CBTF amp is capable of achieving. It has been written in some publications that the individual amps of a push-pull pair do not need to be matched closely. That may or may not be true. I have not expended much effort to confirm or deny that statement. However, for all of my experiments and measurements I matched the individual basic amps as close as possible with $h_{FE}$ of the BJT's matched to within 1 digit using a DVM with

an $h_{FE}$ range, all resistors 2% tolerance or less, the transformers T1 wound as identical as possible and with identical as possible lead lengths, T2 and T3 wound as identical as possible and with as identical as possible lead lengths, ferrite magnetic shielded chokes for RFC1 (Mouser 434-02-102J) to minimize mutual inductance coupling between the individual basic amps (chokes wound on FT-50-43 ferrite toroid cores might have been better for this purpose, but I could not detect any difference between the commercial Mouser chokes and hand-wound toroids in prototype amps which I tested extensively), and PC board construction with the individual basic amps laid out as identical as possible. Perhaps as a result of these precautions and attention to detail, I have been able to construct push-pull CBTF amps with IIP2 of about +100 dBm, which is substantially higher than has been reported for any previous amp, and especially for an amp with 12 dB power gain.

Because so little information was available regarding biasing, the relationship between IIP2 and collector current, the relationship between IIP3 and collector current, the relationship between IIP2 and frequency, the relationship between IIP3 and frequency, and so on, I made extensive studies of these issues and relationships. The results of those measurements are given below in Fig. 3.
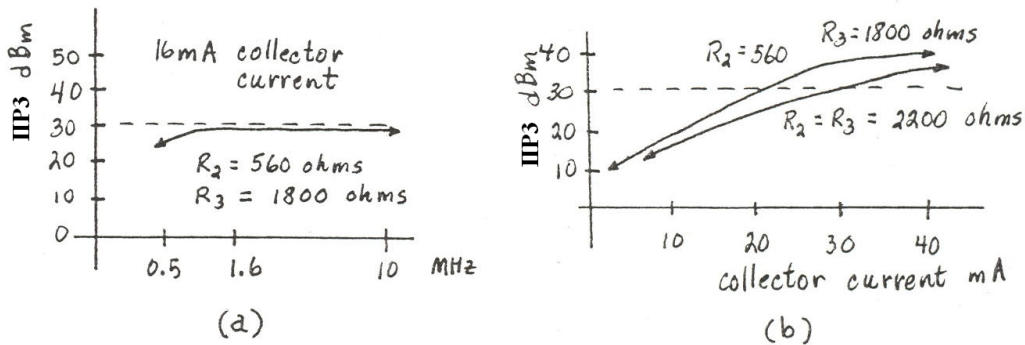
Fig. 3



(a)

(b)

Fig. 3(a) shows the tendency for IIP3 to decrease as frequency decreases within and below the MW band. The biasing used there, R2 = 560 ohms and R3 = 1800 ohms, was the first biasing I tried. With other biasing, the relationship is different, but similar. Fig. 3(b) shows a feature of biasing: the ratio of R2 to R1 should be kept as low as practical to obtain maximal IIP3 with minimal collector current. R2 = 1000 ohms and R3 = 4700 ohms, which was recommended above for the basic amp of Fig. 2, provides a ratio of about 1/5, about the smallest practical ratio. For smaller ratios, adjustment of collector current via R1 becomes progressively more difficult. Also, smaller R2/R3 ratios would likely cause thermal instability of IIP2 for the push-pull amp, and make maximum IIP2 via adjustment of R1 difficult, if not impossible, to obtain.
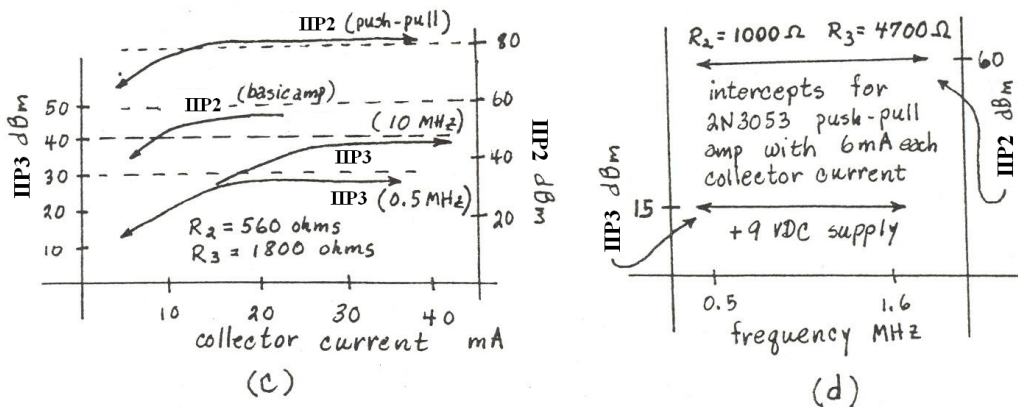


(c)

(d)

Fig. 3(c) compares the relationships between IIP (2 and 3, push-pull and basic amps) and collector current. It should be observed that for a given SW frequency (10 MHz in this case), a "knee" occurs in the IIP curves, and that for greater collector currents beyond this "knee," little or no increase in IIP is obtained. But this is not the case for IIP2 at lower frequencies; see Fig. 3(g) and Fig. 3(h). Fig. 3(d) shows that much higher IIP2 can be obtained in the MW band with a push-pull CBTF amp drawing very low collector current (6 mA in this case) than has been reported for any other kind of amp. Such an amp might be ideal as a two foot air core balanced MW loop amp. The point I want to make here is that the +100 dBm IIP2 which can be achieved with a push-pull CBTF amp is "overkill" for a two foot air core loop amp (unless you want to DX in the shadow of a 50 KW transmitting tower), and that lower collector currents (exactly how much lower, I don't know) would be entirely adequate for such an application.
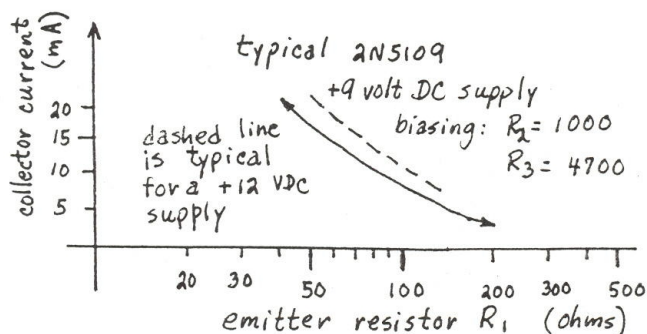


(e)
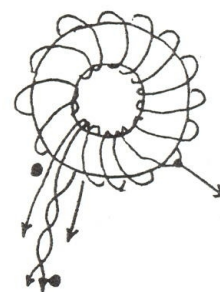


(f)

Fig. 3(e) shows how collector current varies with R1 value for a typical 2N5109 BJT when R2 = 1000 ohms and R3 = 4700 ohms. The purpose of this graph is to give you a starting point for adjusting collector current. Fig. 3(f) is a sketch of T1, a negative feedback transformer wound on a toroid. As mentioned previously, the spacing of the turns is uniform around the entire circumference of the toroid. To get a snug fit, and to preclude the one turn feedback link from making physical contact with the semiconductor material of the 75 material toroid, it is helpful to use a short length of insulation over the wire of the one turn link.
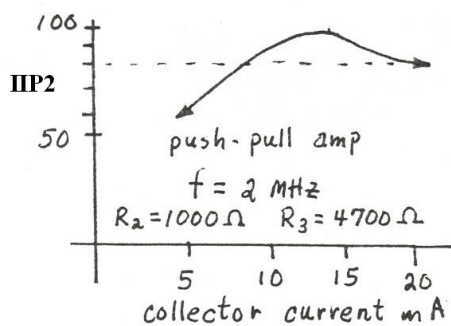


(g)
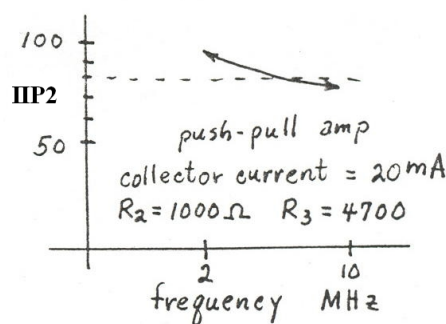


(h)

All of the graphs of Fig. 3 were developed for CBTF amps using a 1:11:4 feedback transformer, i.e., for 12 dB power gain CBTF amps. Similar results may be obtained for CBTF amps with other turns ratios, but I have not made exhaustive studies of those cases, merely some spot measurements to determine if the results for those

other cases were similar (they were).

A push-pull CBTF amp requires somewhat more careful implementation than the basic amp. Nevertheless, the reward is well worth the effort: complete elimination of 2nd order intermodulation distortion products if implemented and used correctly. With the aid of a DVM, you should be able to obtain the required 2% or closer tolerance resistors from a Radio Shack 271-312A package  Since you probably won't have an intermodulation distortion measurement system . of measuring 2nd order intercepts of +80 dBm, much less +100 dBm, you probably won't be able to adjust a push-pull CBTF amp for maximum IIP2. However, you will still have an amp with intercepts of about +80 dBm merely by matching the resistors and 2N5109's as described above. If you will order some 1% tolerance 10 ohm 1/4 watt resistors from Mouser (or your favorite supplier), you can probably do even better by trying different resistors for one and/or the other R1 resistors of the basic amp pair until you have made the collector currents of the two basic amps as nearly equal as possible, where collector currents are measured by measuring the voltages across the two 1% tolerance 10 ohm resistors R4 in each of the basic amps. Or you can simplify this adjustment by using the 25 turn 100 ohm cermet pots recommended previously for the basic amp of Fig. 2. In that case, one pot is set for approximately the current drain desired (again, current drain is measured by measuring the voltage across the 10 ohm 1% tolerance resistor R4 for that amp), and then adjusting the other pot until both collector currents are equal (as indicated by equal voltages across both resistors R4). If you do have an intermodulation distortion measurement system capable of measuring 2nd order intercepts in excess of +100 dBm (which is unlikely), then you merely adjust the collector current of one of the pairs to whatever value is desired, and while observing a 2nd order product produced by the CBTF amp, adjust the second pot to minimize the 2nd order product. Don't even attempt to adjust the push-pull CBTF amp in this manner unless you are certain you know what you are doing. To the best of my knowledge, there is no other intermodulation distortion measurement system than my own which is capable of accurately and reliably measuring 2nd order intercepts in excess of +100 dBm.

The R1 = 100 ohms, R2 = 560 ohms, R3 = 1800 ohms amps discussed above should have collector currents of about 16 mA (32 mA total for the push-pull), and if they do, then they will not need to be heat sinked. If you decide to operate them at higher collector currents (say, to obtain higher IIP3), then they should be heat sinked. Many heat sinks are not easy to use, and some are impossible to use. The Mouser 567-7-120-BA heat sink is one of the easier heat sinks to use, but still difficult. The main problem with many heat sinks for TO-39 cases is that the metal material of the heat sink is not flexible enough to make it easy to mount the heat sink on the TO-39 case. I use small screwdrivers as miniature wedges to slowly open up the Mouser heat sinks until at some point a TO-39 case can be slid into the heat sink with the screwdriver still wedged into the top of the heat sink slot, so that when the wedge (screwdriver) is removed, the heat sink fits tightly (and I do mean tightly). I do not try to wedge the heat sink open in one try, but begin with the smallest possible screwdriver, and move up through progressively larger (but still small) screwdrivers until the condition above is achieved.

You should not pry on the heat sink with a twisting action of the screwdriver. That will cause deep scratches and metal burs on the heat sink, which will make it more difficult to use, and perhaps degrade its heat dissipation characteristics. The Mouser heat sink should be adequate for up to about 30 mA continuously. However, there really is no good reason to run a CBTF amp much above 25 mA because, as shown in Fig. 3, with appropriate biasing, the increase in IIP3 above 25 mA is negligible. And if maximum IIP2 is desired, optimal collector current is in the 13 to 15 mA range, where no heat sink is required.



top view

side view

Fig. 4

I already had several kinds of heat sinks on hand (but not the Mouser 567-7-120-BA) when I started operating CBTF amps at higher collector currents (which require heat sinks). But I did not like any of the heat sinks I had at the time, mainly because they were difficult, if not impossible, to adjust for proper tightness of fit. So initially I made my own heat sinks from 0.021 inch thick copper plate, cut into 11/16 inch wide by 2.5 inch long strips, and bent into the shape shown in Fig. 4 using the shank of a 5/16 inch drill bit. The 0.021 copper plate was obtained from my local sheet metal shop. It is the standard copper plate used to make copper gutters locally. I was given scrap pieces free of charge. They would probably have cut it to the 11/16 by 2.5 size, but I did not know what size I wanted when I got the scrap copper plate. I used a nibbling tool to fabricate

several sizes for testing. The 11/16 by 2.5 size turned out to be sufficient for heat sinking up to about 25 mA continuous collector current, and up to about 40 mA collector current for brief periods. Wider strips may be used for continuous collector currents above 25 mA. Tightness of fit can be adjusted easily with finger pressure: bend the interior circle together until the circle is almost closed, and it should be difficult (but not impossible) to insert a TO-39 case by hand. Removal of a TO-39 case from this heat sink is best done with the aid of a short piece of 3/16 inch or 1/4 inch hardwood dowel. The collectors of 2N5109's are connected directly to the cases, so the cases, and, consequently, the heat sinks, are at +V DC volts. That is one reason why tight heat sink fit is required. The heat sink should not move around, and possibly short the DC supply. The other reason for tight heat sink fit is to provide good thermal contact.
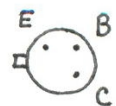
Russell Scotka has been using a push-pull 10.8 dB gain (1:11:4) CBTF amp which I sent him several weeks ago as part of a broadband phased antenna system he has been developing, and I set the amp up for operation with a +12.0 volts DC power supply. Russ' phasing system is based one the one described on page 75 of Victor Misek's *The Beverage Antenna Handbook*, Second Edition. That system uses a VN66AF power FET amp drawing about 90 mA current with a 12 volt DC power supply. With the power FET amp, Russ had several IMD products (all, apparently, 2nd order products). With the push-pull (dual 2N5109) CBTF amp I sent Russ, all IMD products completely disappeared. Russ lives in a high RF urban MW environment in Margate, FL, near Miami, so I doubt there could be a more convincing testament to the effectiveness of a push-pull CBTF amp than the pounding Russ exposed it to. If memory serves me correctly, I set Russ' amp for about 25 mA collector current for each of the basic amps of the push-pull pair, i.e., for a total current drain of about 50 mA. This illustrates an important feature of CBTF amps: you get higher intercepts with rather modest current drains compared to other approaches. Actually, this is not a fair comparison because the power FET amp in Misek's original circuit is not push-pull, and the residual IMD experienced with the power FET amp is, apparently, 2nd order. However, a push-pull version of Misek's amp would require two power FET amps, with a total current drain of about 180 mA. So a push-pull CBTF amp is clearly a better choice for this application based on current drain considerations alone. Also, it is unknown whether a push-pull VN66AF power FET amp would have similar intercepts to a push-pull CBTF amp with the p-p-p FET amp drawing 180 mA.

A free-hand sketch of the bottom view of a PC board layout for a push-pull CBTF amp is shown below in Fig. 5 below. The sketch was drawn enlarged by 1.5625 so that when Fig. 5 is reduced by 0.64 it will be more-or-less exact size. The sketch in Fig. 5 was also drawn enlarged because it would have been impossible for me to produce an exact size drawing. I produced the PC board with Radio Shack dry-transfers and resist pen. The current dry-transfers available at Radio Shack are virtually useless; they don't stick well, they split, and they tend to wash off while the board is being etched. But if you are determined, like me, you can use them. I should explain certain peculiar features of the PC board layout, namely the unused pads. Some of the unused pads are for paralleling bypass and coupling/blocking capacitors. It is sometimes difficult to obtain monolithic ceramic capacitors with values larger than 0.1 uF, so I allowed for paralleling 0.1 uF capacitors to obtain the required 0.2 uF capacitors for flat gain to 100 kHz. I was also unsure if I should parallel 0.2 and 0.0047 uF capacitors for better bypassing and coupling from 100 kHz to beyond 30 MHz (i.e., to maximize the broadband frequency range).

I also wanted the option of having the collector currents adjustable with Spectrol, 1/2 watt, cermet, vertical, top adjust, 25 turn pots, or fixed (without the Spectrol pots, with fixed 1/4 watt resistors). And finally, I wanted the option of making the input or output push-pull (for example, to interface the push-pull amp directly with a two foot balanced air core loop, or to interface the amp directly with other balanced devices, such as noise reducing antennas, or receivers like the R-390A and HQ-180A with balanced inputs). So you will notice that either the input or output (or both) can be "floated" by removing the ground jumper(s), and that either side of the input and output transformers may be grounded. There are also several dark dots which denote holes drilled in the ground plane for ground pins or chassis tie points, and two for resistor ground points if the 25 turn pots are not used (which enable one to obtain non-standard resistance values by paralleling standard values). The cut-out along the center of the PC board is for a 1.75 inch long by 1.75 inch high rectangle of double sided 1/16 thick PC board which is required as a ground plane barrier between the two basic amps of the push-pull pair. The slot may be cut out with a thin emory grinder attachment of a Dremel tool, or with a miniature hack saw (in the latter case the slot will have to be started from one end, and the ground path at that end should be re-established with a heavy

— all fixed resistors are 1/4 watt

— 100 Ω pot is 25 turn Spectrol, vertical, 1/2 watt, 0.1" spacing Mouser 594-64Y101

— 1mH choke is 0.14" spacing, Mouser 434-02-102J

— 330 μH choke is 0.30" body, axial lead, Mouser 43LS334
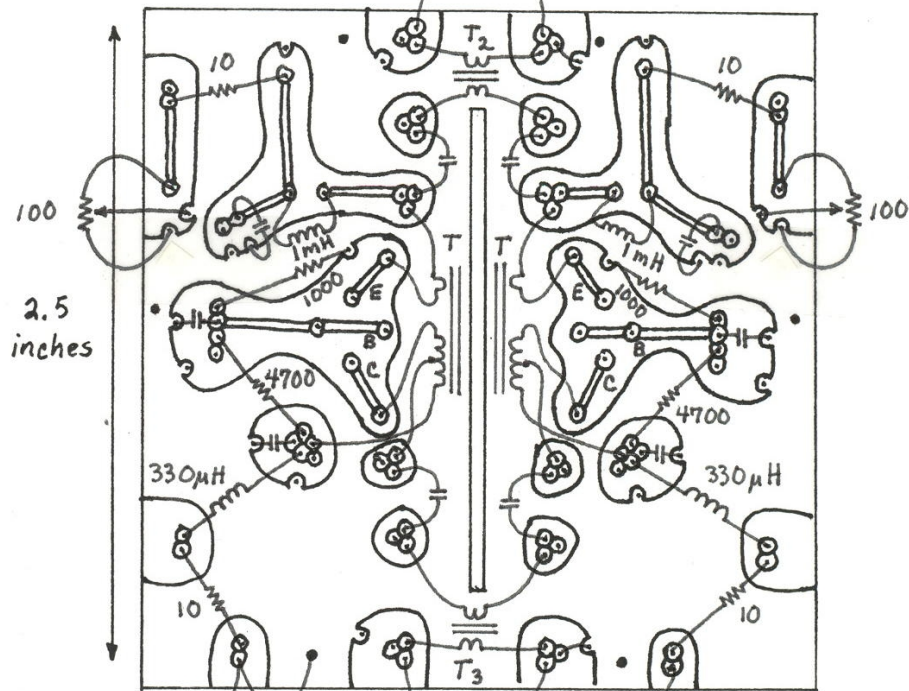
Fig. 5

— the scale above is $1/0.64 = 1.5625$

— the pads and trace widths are not to scale; use larger pads and trace widths

— 2N5109 and 2N3053 pin-out: bottom view

duty jumper).

If you don't like to or don't want to etch a PC board, you may use the Hayward & Hayward "ugly weekender" method to build the push-pull amp. Generally, the "ugly weekender" method involves using high megohm resistors as insulated standoffs (I use 4.7 meg ohm or higher). You can tack-solder the resistors directly to the copper foil of a piece of PC board, as they did, in which case the copper foil will face up towards the components.  Or you can drill holes in the unetched PC board, stick resistor leads through the holes, bend the leads so that the bases of the resistors are flush against the insulated PC board material, and solder the resistor leads, as I do.  The unetched PC board provides an excellent ground plane for the circuit, and circuits constructed by this method work well up into the GHz range when RF is routed with miniature hard line coax.  You will need

a somewhat larger piece of PC board, say 3 inches by 3 inches, and the ground plane barrier between the two basic amps of the push-pull pair will need to be a bit higher, say 3 inches. As a matter of fact, the first push-pull CBTF amp I built with soldered connections used the "ugly weekender" method of construction, and that amp is currently performing flawlessly in Russell Skotka's broadband phased antenna system. I did not use the Spectrol pots in that version, but selected fixed resistor values by hand (using a DVM) to match the collector currents in the two basic amps of the push-pull pair.

At higher frequencies, CBTF amps can be made much smaller using surface mount components because much smaller input, output, and feedback transformers are feasible at higher frequencies. For example, I have built a basic CBTF amp which is about 1 inch square to replace the 45 MHz 1st IF amp in my Drake R8 with IIP3 of about +30 dBm at 45 MHz and about 16 mA collestor current. The amp is a key ingredient in a mod which significantly improves the R8 dynamic range, and is described in my recent article, "Drake R8: Increased Dynamic Range, Mod 2," which should appear soon in *DX News* and *DX Monitor*. It is really not worth the effort to use surface mount components in the 100 kHz - 30 MHz amp described above because the transformers, chokes, and BJT's take up most of the space.

Another application of CBTF amps, already mentioned above, is to broadband phased wire antenna systems. As pointed out above, Victor Misek's phasing circuit is greatly improved when the original power FET amp is replaced with a push-pull CBTF amp. Other phasing approaches which require amps, such as Gerry Thomas' (1985 DX News) "Phase One" delay line circuit (adapted from John Webb's SW circuit described in Oct. 1982 QST, and recently revived by Mark Connelly in his article, "DL-1 Delay Line Phasing Unit," DX News, Vol. 61, No. 26, May 23, 1994), may also provide much improved performance when a push-pull CBTF amp is used. Mark recommended against using an amplifier with a delay line phasing circuit in a high RF urban environment, and mentioned amplifier outputs of +20 dBm for strong local stations. But such high outputs are impossible with a 12 dB nominal gain push-pull CBTF amp, reasonable length wire antennas (say, 100 feet long or less), and a phasing circuit (which introduces loss), assuming that the listening location is a reasonable distance away from a 50 KW transmitting tower. I'd say that if you have -10 dBm on your antenna, then either your antenna is too long, or you live too close to the transmitting tower. My 1 KW super-local KRUS puts about -21 dBm on my wire antennas, so that with a 12 dB nominal gain amp, I am looking at about -9 dBm. However, I would not use the 12 dB nominal gain amp unless I had about 12 dB loss in my phasing system. Broadband amps, even ultralinear broadband amps, should only provide enough gain to make up loss, no more.

Added 1/1/05: This article was originally published in *DX News* in 1994. Curiously, no push-pull Norton amps appeared in consumer grade receivers until just this past year, in the ICOM IC-7800 and IC-756ProIII(3). For the hobbyist, the 2N5109 is still often the BJT of choice if you "roll their own" and if space is not an issue. Not long after my article appeared, Kiwa Electronics started producing push-pull Norton amps using Motorola MRF581A's which are widely used by DXers. At right is a 10.8 dB gain push pull transformer feedback Norton amplifier which uses MiniCircuits T1-6 transformers at the input and output. If you want to make really tiny Norton amps, the 581A's are the BJT's to use. The non-A's can also be used and are equally small. Both are still available through RF Parts, and perhaps other places. They are not cheap. But you can put 25 mA through them without any heat sink, which amazes me. I trim the flanges off their collector leads before I slip on ferrite beads, which leaves just enough
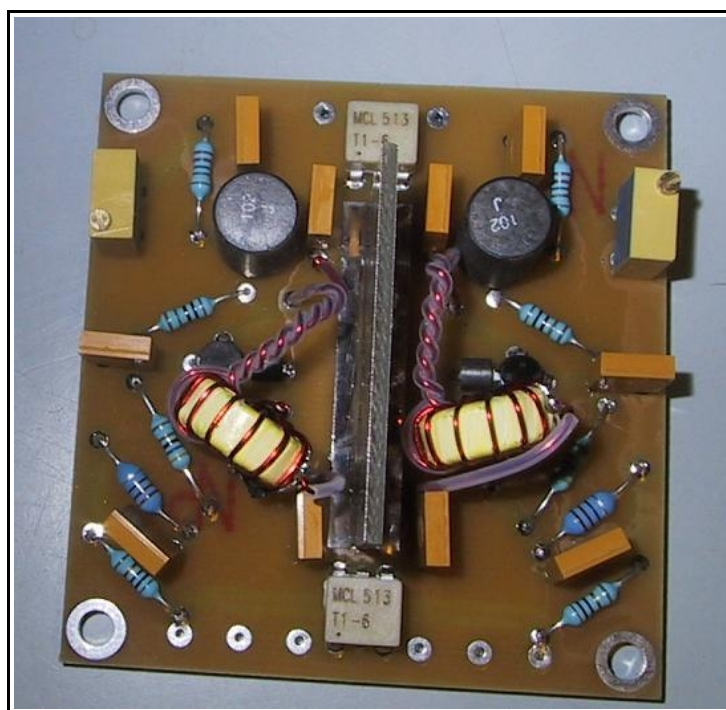
lead protruding from the ferrite bead to solder surface mount or through hole.   For many years I have used R2 / R3 = 1000 / 4700 ohms for the bases voltage dividers instead of the original 560 / 1800 ohms.  The MiniCircuits T1-6 transformers may not be such a good idea.  There have been reports of excessive IMD in other circuits due to MiniCircuits broadband transformers compared to physically larger ferrite toroid transformers.  If you have higher than normal RF levels, you should use Amidon FT-50-70 ferrite toroids as specified earlier in this article.

Added 5/20/07:  While it has been almost 13 years since the original article was written, little has changed.  Push- pull common base transformer feedback amplifiers are still my amplifiers of choice where high intercepts are required, and I still use 2N5109's, and now also MRF581A's which Craig Siegenthaler (KIWA) found.  Radio Shack stores no longer carry 2N3053 BJT's, and some people found them unsatisfactory, so I deleted references to and discussions of them.  Some other parts of the original article have also been deleted or revised.  If you want a copy of my original article, it should still be available as a reprint from The National Radio Club as reprint A87.  Several other BJT's, such as 2N3866's, 2N5943's, and MRF 586's, can be used, but they are generally more expensive, and do not provide better IMD performance or noise figures.