West Indian Journal of Engineering Vol. 23, No. 1, (July 2000) Technical Paper (Gift & Copeland) 28 - 35

A High-Quality Mixer Circuit

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This paper describes a new mixer circuit that uses a true trans-resistance amplifier. As a result of this, the closed-loop bandwidth is independent of the number of signal inputs and channel crosstalk is minimised. The circuit employs an operational amplifier preceded by a single-transistor amplifier operated in the common base mode.

1. Introduction

An electronic mixer enables the linear combination of a number of individual signals into a single composite signal. Such a processing function is used extensively in the recording and broadcasting industries as well as in domestic audio. In order to achieve the high performance demanded in such applications, a mixer must have excellent channel isolation characteristics, in order to minimise inter-channel crosstalk.

In this paper, we discuss the limitations of the conventional mixer circuits based on the inverting operational amplifier (op amp) configuration. Specifically, the crosstalk characteristics and the effect of increasing the number of input channels are highlighted. A new design with improved characteristics is then described and test results obtained from a practical implementation given.

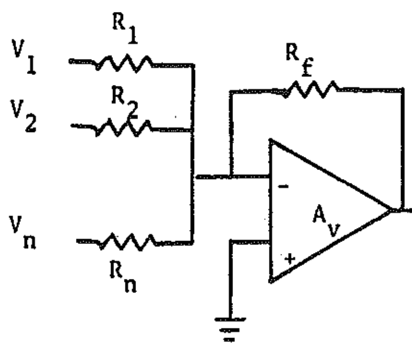
2. Conventional Mixer Circuit

The conventional mixer circuit employs an operational amplifier in the summing inverter configuration (**Figure 1**). Here, A_v is the open-loop voltage gain of the op amp which is very large. Negative feedback around the amplifier reduces the normally high input impedance at the inverting input and there creates a virtual earth. This node serves as the summing junction for the various input signal currents. To satisfy stability requirements, the amount of negative feedback must decrease with increasing frequency and hence the closed-loop input impedance at the inverting input rises with increasing frequency. The quality of the virtual earth therefore deteriorates and this results in increased inter-channel crosstalk in the upper audio frequency spectrum.

To further demonstrate this, consider the mixer circuit of **Figure 2**. A signal voltage V_1 with a source resistance R_{S1} is applied to the circuit via an input resistance R_1. The voltage V_o appearing across R_{S2} (corresponding to the source resistance of the input signal of this second channel) is the crosstalk voltage due to inadequate channel isolation.

Using the equivalent circuit of **Figure 2**, the voltage V at the inverting node is given by:-

$$V = \frac{R_{in} // \left(R_2 + R_{s_2}\right)}{R_1 + R_{s_1} + R_{in} // \left(R_2 + R_{s_2}\right)} V_s \qquad (1)$$



$$V_o = -R_f\left[\frac{V_1}{R_1} + \frac{V_2}{R_2} + \dots \frac{V_n}{R_n}\right]$$

FIGURE 1: *Conventional Mixer Circuit*

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Pertinent discussion will be published in January, 2002 West Indian Journal of Engineering if received by November, 2001.

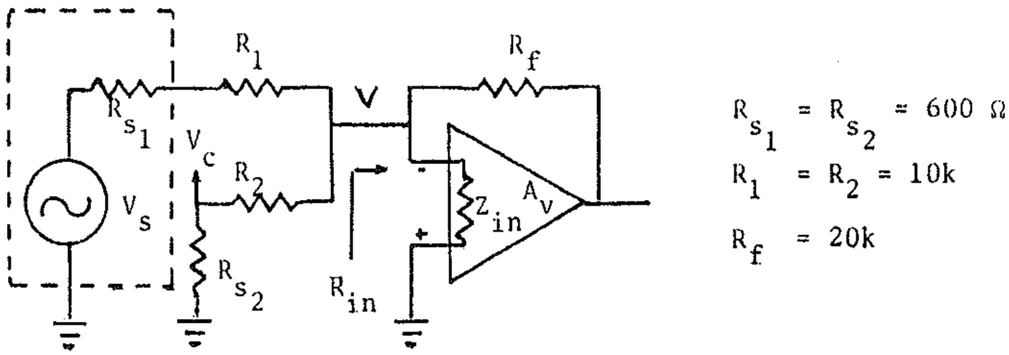

FIGURE 2: *Measurement of Inter-Channel Crosstalk for Conventional Mixer*

where from Miller's Theorem [1][1]

$$R_{in} = \frac{R_f}{A_v + 1} \simeq \frac{R_f}{A_v} \quad A_V >> 1 \text{ and } Z_{in} \text{ large} \qquad (2)$$

Hence the crosstalk voltage V_c is given by

$$V_c = \frac{R_{s_2}}{(R_2 + R_{s_2})} V \qquad (3)$$

which after substituting (1) and re-arranging yields

$$\frac{V_c}{V_s} = \frac{R_{s_2}}{(R_2 + R_{s_2})(R_1 + R_{s_1})}$$

$$\frac{R_{in}}{1 + R_{in}\left(\frac{1}{R_2 + R_{s_2}} + \frac{1}{R_1 + R_{s_1}}\right)} \qquad (4)$$

Assuming $R_2 >> R_{s_2}$ and $R_1 >> R_{s_1}$, (4) reduces to

$$\frac{V_c}{V_s} = \frac{R_{s_2}}{R_1 R_2} \frac{R_{in}}{1 + R_{in}\left(\frac{1}{R_1} + \frac{1}{R_2}\right)} \qquad (5)$$

Because general purpose operational amplifiers normally have the dominant pole at a frequency close to DC, A_v is modelled as an integrator

$$A_v(s) = \frac{\omega_o}{s} \qquad (6)$$

where $\omega_o = 2\pi f_o$ (7)

and f_o is the unity gain bandwidth of the operational amplifier. R_{in} in (2) therefore becomes

$$R_{in} = \frac{R_f}{\omega_o} s \qquad (8)$$

which when substituted in (5) yields

$$\frac{V_c}{V_s} = \frac{\left(R_{s_2} R_f' / R_1 R_2\right)s}{1 + \left(R_f' / R_T\right)s} \qquad (9)$$

where $R_f' = R_f / \omega_o$ (10)

and $\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2}$ (11)

[1] Jung (Ref. [2] p. 169) incorrectly gives $R_{in} = \frac{R_f}{A_v \beta}$ where β is the feedback ratio.

For the circuit of **Figure 2**

$$R_1 = R_2 = 10k,\ R_f = 20k,\ R_{s_2} = 600\Omega$$

and f_o=1M Hz. Hence (9) becomes

$$\frac{V_c}{V_s} = \frac{\frac{6}{10^8 \pi} s}{1 + \frac{2}{10^6 \pi} s}$$

Experimental evaluation of the circuit was performed using V_s = 1 volt over a frequency range 10 kHz to 1 MHz. (Crosstalk voltage for frequencies below 10 kHz were unmeasurable with the test equipment available). The test results and the theoretical curve are plotted in **Figure 3**. The two curves show a close correlation at the lower frequencies but deviate at the higher frequencies. This is due to the falling off of the operational amplifier open-loop input impedance Z_{in} at the higher frequencies. The assumption made in (2) that Z_{in} is large therefore becomes progressively invalid. The theoretical curve can therefore be viewed as a worst case situation.

A further problem exists with the conventional mixer circuit. Because Z_{in} is normally high, the circuit functions essentially as a voltage amplifier (i.e., voltage sampling/ voltage feedback) and not as a transresistance amplifier (i.e., voltage sampling/current feedback) - the signal feedback to the inverting input is attentuated according to the equivalent resistance of the input summing resistor network. Consider the circuit of **Figure 4**, where it is assumed that each input channel resistor is fed from a low source impedance.

$$\text{Let } R = R_1 // R_2 // ... // R_n \tag{12}$$

$$\text{and} \quad R_1 = R_2 = ... R_n = R_f \tag{13}$$

$$\text{Then} \quad R = R_f / n \tag{14}$$

and the feedback ratio β is given by

$$\beta = \frac{R}{R + R_f} = \frac{R_f / n}{R_f / n + R_f} = \frac{1}{1+n} \tag{15}$$

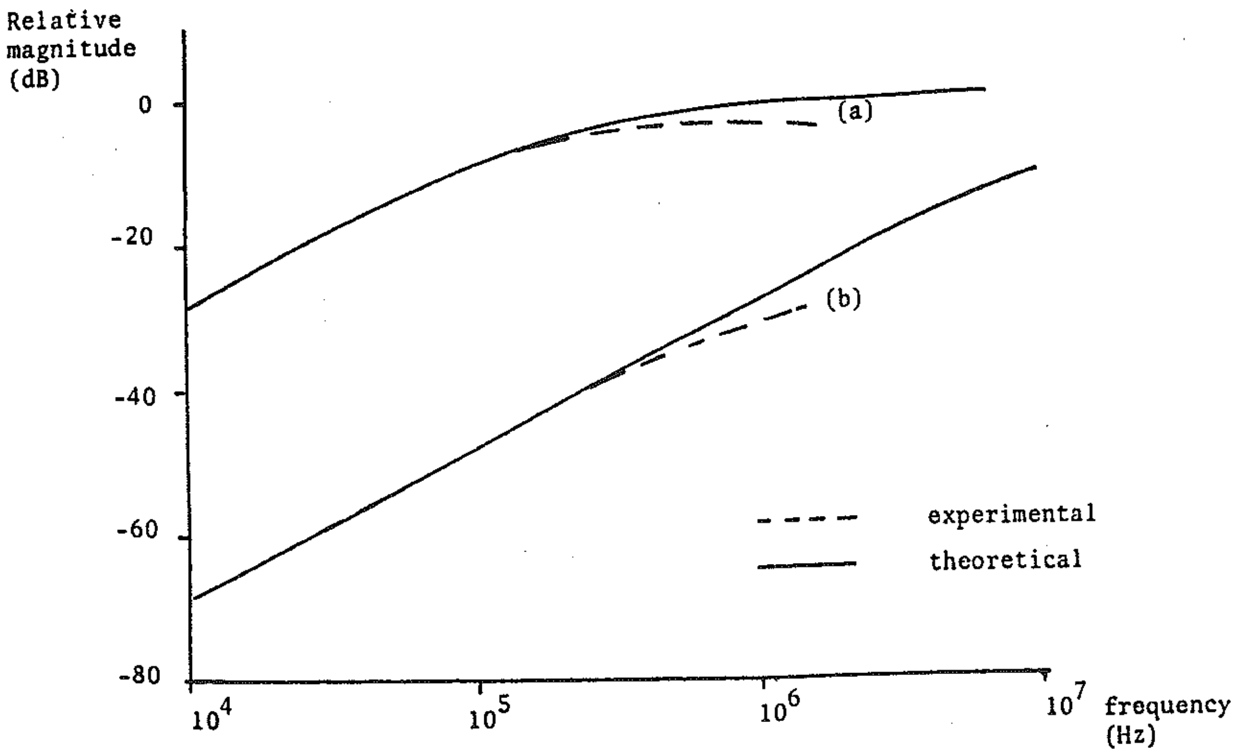

FIGURE 3: *Theoretical and Experimental Crosstalk Characteristics for (a) Conventional Mixer Current and (b) New Mixer Circuit*

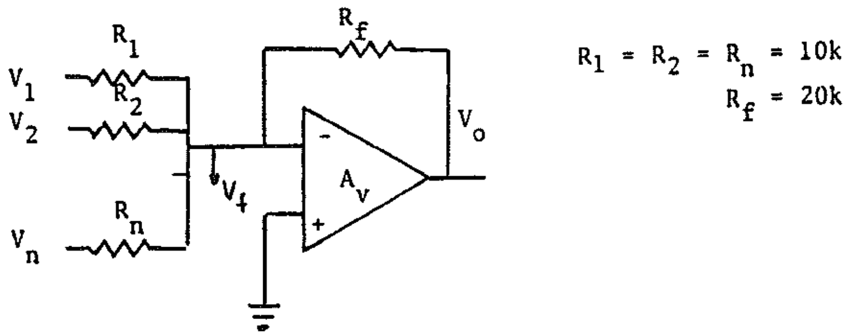

FIGURE 4: *Measurement of Bandwidth Reduction for Conventional Mixer Network*

where β is defined by

$$\beta = \frac{V_f}{V_o} \tag{16}$$

The noise gain is given by

$$\frac{1}{\beta} = 1 + n \tag{17}$$

The effective circuit closed-loop transfer function (voltage gain) is given by

$$A_{v_{cl}}(s) = \frac{-A_v(s)}{1+\beta A_v(s)} \frac{R_f}{R+R_f}$$

$$= \frac{-1/\beta}{1+\left(1/\beta\omega_o\right)s} \cdot \frac{R_f}{R+R_f} \tag{18}$$

Hence the closed-loop bandwidth f_{cl} is given by

$$f_{cl} = \frac{\beta\omega_o}{2\pi} = \beta f_o = \left(\frac{1}{1+n}\right) f_o \tag{19}$$

For n>>1, $f_{cl} \cong \frac{f_o}{n}$ (20)

Thus, although the gain of each individual channel is constant, with the increase of input channels, from (15) the amount of feedback decreases and equivalently from (17), the noise gain increases. This manifests itself as a decrease in closed-loop bandwidth as indicated in (20). Harmonic distortion will also increase.

The reduction of system bandwidth with increase in n was experimentally verified using the circuit of **Figure 4**. Channel 1 was driven by a low impedance signal source with the (n-1) remaining input channel resistors connected to ground. The resulting bandwidth measurements for various values of n are shown in **Figure 5**.

3. New Mixer Circuit

The new mixer circuit shown in **Figure 6** overcomes both of the disadvantages of the conventional mixer circuit by employing a common base transistor amplifier at the input of the operational amplifier and inside the feedback loop. This additional stage provides the operational amplifier with a low open-loop input impedance at its inverting input, thereby making it ideal for the summing of input signal currents even before feedback is applied. Negative feedback around the hybrid arrangement further enhances the already low input impedance.

Because of the low open-loop impedance, the input signal can be viewed as a current, thereby making the circuit a true transresistance amplifier. The open-loop transresistance is given by

$$R_m = \frac{V_{out}}{I_{in}} = -A_v(s)R_L = R_L\frac{-\omega_o}{s} \tag{21}$$

The closed-loop transresistance is therefore

$$R_{m_{cl}} = \frac{R_m}{1+\beta R_m} \tag{22}$$

where for this circuit,

$$\beta = \frac{1}{R_f} \tag{23}$$

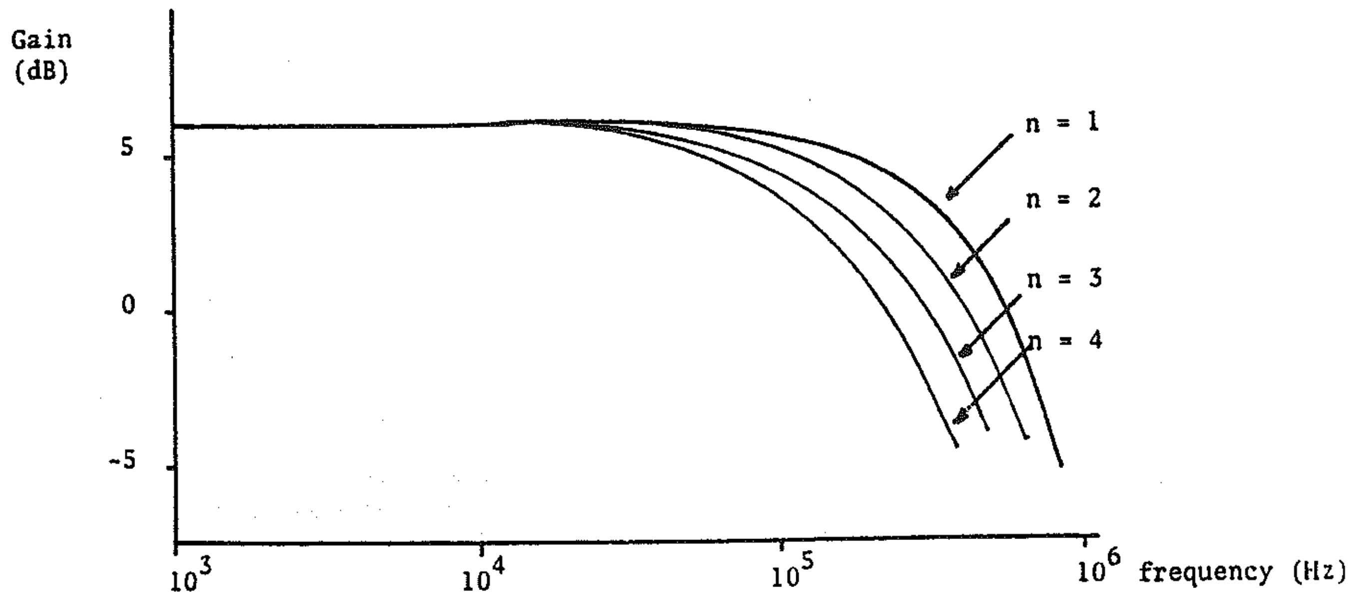

FIGURE 5: *Frequency Response Variation with n for Conventional Network*

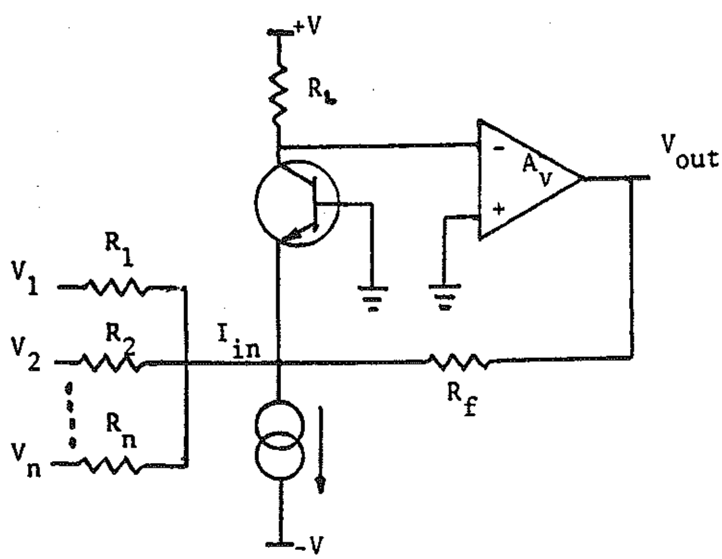

FIGURE 6: *New Mixer Circuit*

(giving $R_{m_{cl}} = R_f // R_m$)

Re-arranging,

$$R_{m_{cl}} = \frac{-R_f}{1 + \frac{R_f}{R_L \omega_o} s} \tag{24}$$

The closed-loop bandwidth f_o' is therefore given by

$$f_o' = \frac{1}{2\pi T_o'} \tag{25}$$

where $T_o' = \dfrac{R_f}{R_L \omega_o}$ (26).

i.e., $f_o' = \dfrac{R_L}{R_f} f_o$ (27)

In a transresistance amplifier, because the amount of negative feedback that can be applied is not limited as in the case of a voltage amplifier, the closed-loop bandwidth f_o' can in principle be increased beyond f_o' by making $R_f < R_L$. Parasitic system poles in practice set an upper limit on f_o'. The important point here, however, is that f_o' is independent of the number of channel inputs since the feedback ratio is set by R_f.

The voltage gain of a single channel is given by

$$\frac{V_{out}}{V_{in}} = \frac{R_{m_{cl}}}{R_1}$$

Thus the value of R_f also influences the channel gain. For the transresistance amplifier, the closed-loop input impedance R_{in} at the emitter of the input transistor is given by

$$R_{in} = \frac{R_{cb}}{loop.\ gain} \tag{29}$$

where R_{cb} is the common base input impedance

$$R_{cb} = \frac{m}{40 I_c} \tag{30}$$

with I_c being the standing collector current and $m \simeq 1$. The loop gain is given by

$$loop\ gain = R_m \beta = \frac{R_L}{R_f} \cdot \frac{\omega_o}{s} \tag{31}$$

Hence, using equation (5), the crosstalk for the new circuit is given by

$$\frac{V_c}{V_s} = \frac{R_{ch} / loop\ gain}{1 + (R_{ch} / loop\ gain)(1/R_1 + 1/R_2)} \cdot \frac{R_{s_1}}{R_1 R_2} \tag{32}$$

which after substituting for the open loop gain and re-arranging becomes

$$\frac{V_c}{V_s} = \frac{(R_{s_2} R_f' / R_1 R_2)\rho s}{1 + (R_f' / R_T)\rho s} \tag{33}$$

where $\rho = R_{cb} / R_L$ (34)

Note that since normally $R_{cb} << R_L$, then $\rho << 1$. Thus, comparing (33) with (9), it can be seen that the crosstalk of the new circuit is reduced by factor $\rho << 1$ as compared to the conventional circuit for frequencies below

$$f_p = \frac{R_T}{2\pi R_f' \rho} \tag{35}$$

Above this frequency, the crosstalk performance for both circuits become similar.

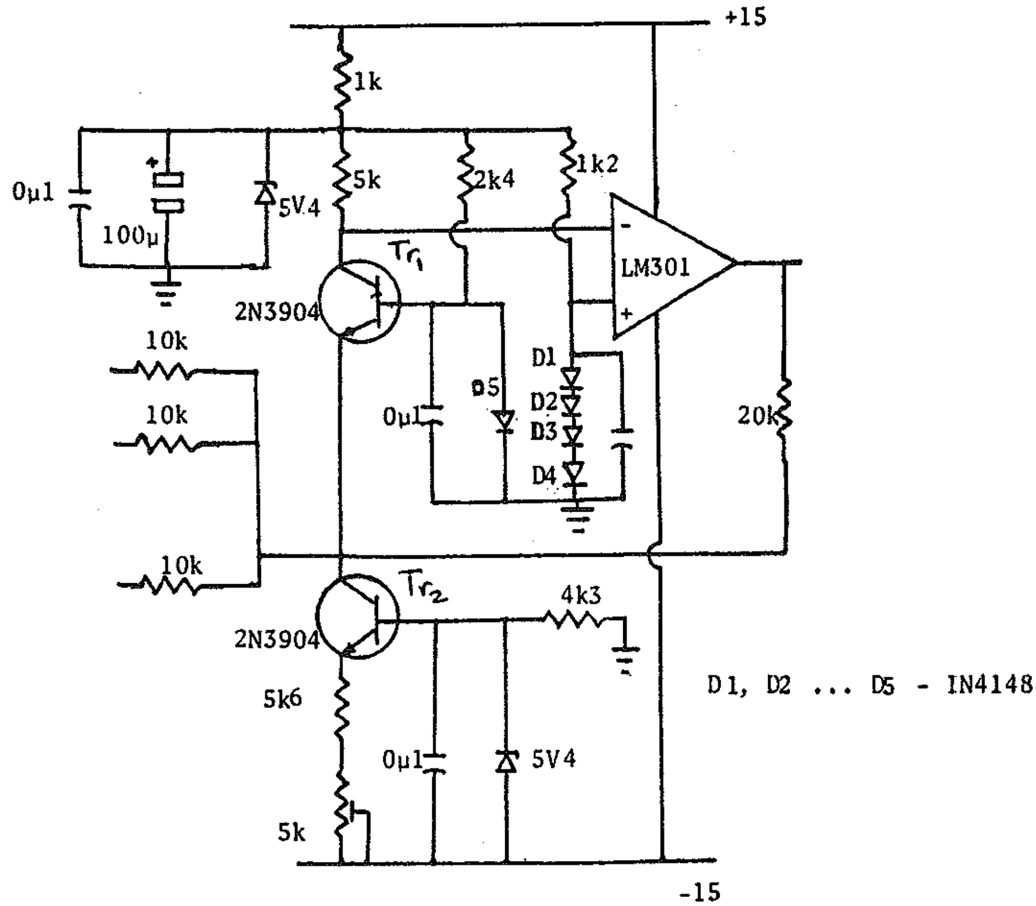

FIGURE 7: *Practical Realisation of New Mixer Circuit*

4. Practical Circuit

A practical implementation of the new circuit is shown in **Figure 7**. Transistor Tr_1 provides the common base input to the operational amplifier while Transistor Tr_2 is a constant current source. Diode D_1 sets the base of Tr_1 at about 0.6 volts so that the emitter of Tr_1 is close to ground potential. The non-inverting input of the operational amplifier is also provided with an offset voltage so that the collector of Tr_1 can accommodate a reasonable voltage swing. The value of R_L is limited by the collector/base capacitance Tr_1, with which it forms a parasitic pole. R_L must be chosen so that this pole is made non-dominant. A value of 5k provided excellent results. The collector current Tr_1 is set at 0.5 mA and this sets R_{cb} at

$$R_{cb} = \frac{m}{40I_c} = \frac{1}{40x0.5x10^{-3}} = 50\Omega$$

This yields $\rho = \frac{R_{cb}}{R_L} = \frac{50}{5k} = 0.01$

Using the same values for R_1, R_2, R_f and R_{s_2} with $f_o = 1$ M Hz, (33) becomes

$$\frac{V_c}{V_s} = \frac{\frac{6}{10^{10}\pi}s}{1+\frac{2}{10^8\pi}s}$$

and $f_p = 25\ MHz$

$$f_o' = 250\ kHz$$

The circuit was tested both for crosstalk figure and bandwidth stability with varying input channel number. The results of the crosstalk test are indicated on **Figure 3** which establishes the improvement of the new circuit as compared to the old. Also, the circuit bandwidth f_o' remained constant at 250 kHz for values of n from 1 to 10.

5. Conclusion

In this paper, the crosstalk limitations and the reducing bandwidth arising from increasing channel input number for the conventional op amp mixer are discussed and demonstrated. A new circuit with improved performance was then presented. Its bandwidth is independent of the number of input channels and the crosstalk is two orders of magnitude less than that for the conventional circuit.

The basic principle embodied in the new circuit converts the standard op amp from a voltage amplifier to a transresistance amplifier and can be applied to the wide range of op amps currently available. Recently, integrated circuit transresistance amplifiers have become available (Ref. [3]) though these are comparatively few in number. In light of the results in this paper, we are currently examining the performance of samples of these ICs in a mixer configuration and will report on our findings in a future paper.

References

[1] Millman, J. and Halkias, C.C. (1972). "Integrated Electronics". McGraw Hill, New York.

[2] Jung, W.G. (1981). "Audio IC Op-Amp Applications". Howard W. Sams, Indiana, USA.

[3] Ogden, F. (1992). "The Current Alternative to Operational Amplifiers". Electronics World & Wireless World. pp. 643.