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Integrated microwave circuits for digital-to-analogue converters based on high-temperature superconducting Josephson junction arrays

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Abstract. A digital-to-analogue converter based on series arrays of high-temperature superconducting Josephson junctions can work properly when the microwave bias current distribution in large arrays is uniform. Unfortunately, the considerable microwave attenuation per overdamped junction with normal resistances $R_N \approx 0.1\ \Omega$ results in a strong total attenuation of the microwave signal in a traditional layout. Also, the meander geometry of the array, necessary when using bicrystal junctions, disturbs the uniformity of the rf current, especially in the millimetre waveband. To overcome these drawbacks we coupled the meander array inductively to a parallel transmission line. The mechanism of microwave coupling in which both external and mutual frequency locking takes place was studied numerically. A circuit with a coplanar waveguide feed line was fabricated and successfully tested at an operating frequency of 30 GHz. A microstrip line with a small attenuation for pumping of junction arrays at frequencies up to 120 GHz was demonstrated.

1. Introduction

Recently a new concept of a digital-to-analogue converter (DAC) based on series arrays of high-temperature superconducting (HTS) Josephson junctions with nonhysteretic current–voltage characteristics (IVCs) was suggested [1]. The DAC can work properly when the ac bias current distribution in large arrays is uniform. HTS Josephson junctions demonstrated characteristic voltage $V_c \approx 200$–$300\ \mu$V at liquid nitrogen temperatures and are very attractive for using DAC at the millimetre wave band microwaves bias current. It can operate at the same microwave frequencies as conventional Josephson voltage standard systems based on niobium tunnel junction arrays. Unfortunately, the considerable microwave attenuation per overdamped junction results in a strong total attenuation of the microwave signal. Also currently available substrate materials, suitable for fabrication of HTS junction arrays, exhibit too large relative dielectric permittivity $\varepsilon > 10$ and dielectric losses $\tan\delta > 10^{-3}$, when applying at frequencies about 100 GHz. Additionally, the meander geometry of the array, necessary when using bicrystal junctions, disturbs the uniformity of the microwave current, especially in the millimetre waveband. To overcome these drawbacks we have suggested coupling the meander array inductively to a parallel transmission line [2]. In this paper we will discuss in detail the mechanism of microwave coupling in which both external and mutual frequency locking takes place. The microwave properties of a microstrip feed line on a thin substrate were studied.

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2. Concept of the integrated microwave circuit

The discussed microwave circuit consists of two basic parts: the Josephson junction array and the microwave transmission line (figure 1). In our experiments we have used arrays made by meandering Au–$YBa_2Cu_3O_7$ bilayer strips across the grain boundary in yttria-stabilized-zirconia (YSZ) bicrystal substrates [2]. For dc current the junctions are connected in series. For ac current the array consists of loops connected in parallel. Each loop includes two Josephson junctions, inductance L and two capacitors C (figure 1). This design supports

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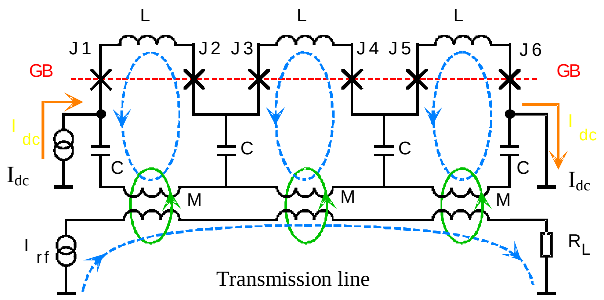

Figure 1. Simplified scheme of the integrated microwave circuit.

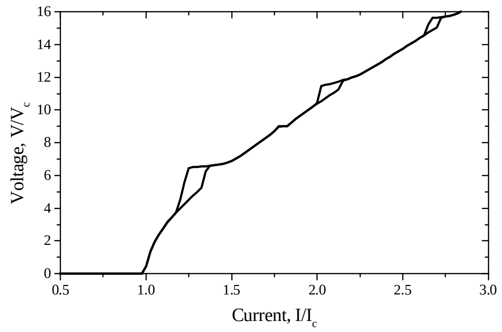

Figure 2. Current–voltage characteristic for a six-junction array with $Z_L/Z_C = 2$ under the influence of an external ac-current source at $f = 1.5\ f_c$, $I_c = 1$ mA, $V_c = 100\ \mu$V.

both the mutual phase locking of the junctions in neighbouring loops and the inductive coupling to the transmission line.

A standard microstrip line (MSL), a coplanar waveguide (CPW) or a slot line can be used as feed lines. This provides feeding of microwave power to each two-junction loop and induces rf bias current in it. The resistive matching load R_L is important for obtaining an uniform rf distribution. It prevents reflections that would create a standing-wave pattern in the array. The decay of the driving rf amplitude along the loop line, which appears due to the dissipation of the microwave power in the stripline, can be easily compensated by an appropriate variation of mutual inductances M by geometrical changes in the design.

The concept developed allows us to choose the substrate materials for the junction array and the transmission line independently, and thus to design circuits with parameters close to the optimum values.

3. Design simulation and fabrication of the microwave circuits

Two different types of circuits have been designed and fabricated. They were optimized for drive frequencies in the interval from 20 GHz to 40 GHz and from 70 GHz to 120 GHz respectively. CPW and MSL were used as feed lines in low and high frequency designs respectively. Optimal parameters for the circuit design had been found previously by simulating device operation.

3.1. Simulations

It follows from figure 1 that the character of the Josephson junction microwave connection depends on the relation between inductive Z_L and capacitive Z_C impedances at the working frequency. This is supposed to be a factor of 1.5–2 larger than the characteristic frequency f_c of the individual junctions. At $Z_L \ll Z_C$ all the junctions in the structure are connected in series and their mutual synchronization is very difficult to achieve. However, at $Z_L \gg Z_C$ the loops with the junctions are connected in parallel, which leads to the desired effects. The results of numerical simulations by means of the PSCAN program [3] confirm this simple analysis. The simulated circuit consisted of a six-junction array. A 5% spread of the critical currents and normal junction resistance was taken into account. In this case the stability of a given in-phase solution can be proved. It was shown that for $Z_L/Z_C = 0.2$ no phase locking would occur. The current–voltage characteristics of six junctions in the array in the case $Z_L/Z_C = 2$ are shown in figure 2. The resonance frequencies of the two feedback loops can be seen clearly. The first peculiarity near the voltage $V \approx 6.9\ V_c$ corresponds to the resonance frequency $f = 1.15\ f_c$ of the large feedback loop (all three inductances L and two capacitances C, see figure 1). The second peculiarity near the voltage $V \approx 12\ V_c$ corresponds to the frequency $f = 2.0\ f_c$ of the small feedback loop.

Applying microwave irradiation at a frequency 1.5 f_c leads to the formation of the current step. We can further enlarge the current step height by increasing the amplitude of the external frequency source. The calculation confirms that at the same time the regions of mutual coherence are destroyed by competition from an external locking mechanism.

3.2. Circuit with CPW feed line

Figure 3 shows the layout of the integrated microwave circuit with CPW feed line (3) [2]. The meander line (1) runs along the grain boundary, representing the series array of shunted bicrystal Josephson junctions, 256 in total. One side of each meander loop is connected to the strip line (2) with the area 0.025 mm^2. The 1 μm thick SiO_2 film was deposited on the top of the strips to provide the isolation between the superconductor and the outer conductor (3b) of the CPW (3). The central (3a) and outer (3b) conductors of the CPW were fabricated from 0.7 μm thick Au films.

The values of inductive and capacitive impedances estimated from the layout geometry ($Z_L/Z_C \approx 1.5$) were close to the parameters which had been used in the simulation. Under external irradiation in the frequency range of 30 GHz the first current step was demonstrated on eight different segments of the array, containing two, two, four, eight, 16, 32, 64 and 128 junctions [2].

Nevertheless, even at these relatively low rf frequencies, the main contribution to the attenuation in the CPW is determined by the dielectric loss $\tan\delta$ in the YSZ substrates. From the measured attenuation at a temperature of 78 K [2], it can be shown that $\tan\delta$ increases in the frequency range from 10 GHz to 40 GHz as $\tan\delta \approx 1.05\times10^{-3} f\ (\mathrm{GHz}) - 2\times10^{-2}$. Extrapolating this dependence to $f = 100$ GHz we obtain

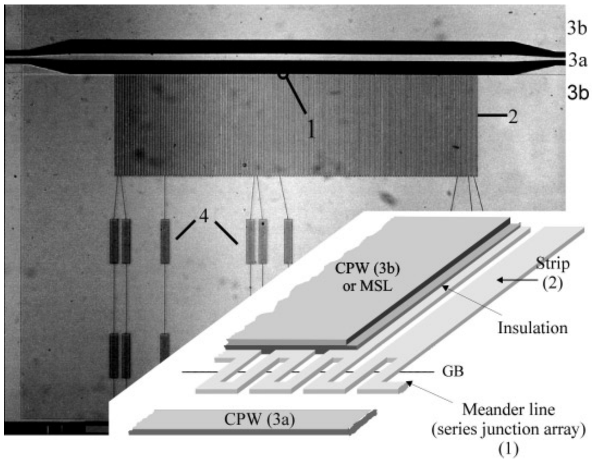

Figure 3. Schematic layout of the new microwave circuit. The CPW is partly lifted to show the insulation layer and the microstrips below. Background: top view photograph.

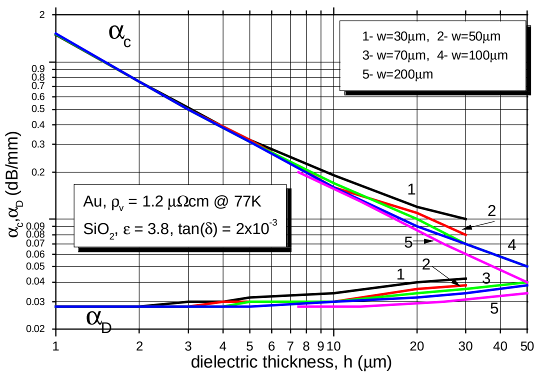

Figure 4. Dependencies of MSL conductor α_c and dielectric α_D losses on dielectric thickness h.

$\tan\delta \approx 8.5 \times 10^{-2}$, which leads to considerable attenuation $\sim$3 dB mm^{-1}. It is for this reason that in our high frequency version of the circuit we focus on the MSL feed line.

3.3. MSL feed line

We supposed that this MSL should be fabricated on a flip chip and would consist of thin gold films deposited on both sides of a thin dielectric substrate. Figure 4 shows the attenuation in the MSL at a frequency of 75 GHz due to losses in the conductor (Au) and dielectric (SiO_2) as a function of dielectric thickness for varying microstrip width w. It is clearly seen that at a thickness larger than 20 μm the losses in the conductor are greatly suppressed, while the dielectric losses remain small.

For the experimental testing a special circuit, consisting of the MSL with two finline taper antennas [4] connected by a microstrip line with two bends, was fabricated. Mica with a dielectric permittivity of 6–6.5 and the thickness of 25 μm was used as a substrate. The width and length of the central strip was 0.13 mm and 8 mm respectively. The wave impedance of the MSL was equal to 16 Ω.

The measured values of the MSL attenuation and the voltage standing wave ratio (VSWR) are shown in figure 5

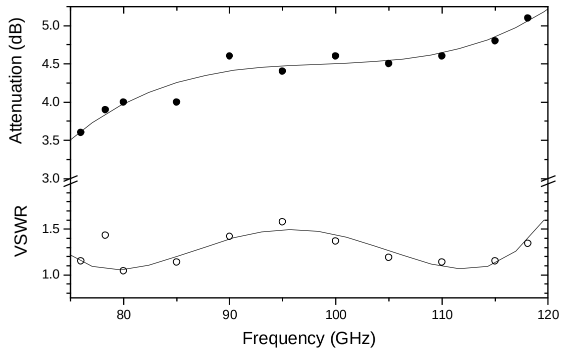

Figure 5. Attenuation (filled circle) and VSWR (unfilled circle) for 8 mm length MSL with two finline antennas. Lines are drawn to guide the eye.

as a function of microwave frequency from 75 GHz to 120 GHz. The data demonstrated good matching between a WR-10 waveguide and the microstrip via the two finline taper antennas. The SWR does not exceed 1.6 throughout the whole frequency band. Full attenuation at the frequencies from 70 GHz to 90 GHz, typical for the classical voltage standard, does not exceed 4.5 dB. Taking into account that the main contribution to the attenuation comes from the two finline taper antennas, we can conclude that the real attenuation in the line is of the order of 0.1 dB mm^{-1}. This value is close to that of the design requirement.

4. Summary

The new method proposed here for coupling external microwave irradiation with the Josephson junction array offers a solution to one of the main problems of the DAC device. Namely, we can combine the advantage of larger junction characteristic frequencies (provided by their resistance $\sim$0.1 Ω) and simultaneously get rid of the large rf losses, which occurred in the case of traditional layout solutions. Correspondence between measurements at a frequency of 30 GHz and the results of simulation confirm the validity of the developed approach. The fabricated flip chip MSL, with low attenuation in a wide frequency band, provides an opportunity for a further increase of the irradiation frequency of HTS Josephson junction arrays for DAC.

Acknowledgments

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